B.R. 230

DMIRALTY HANDBOOK OF WIRELESS TELEGRAPHY

Volume II

Wireless Telegraphy Theory

March, 1938

The Lords Commissioners of the Admiralty have decided that a standard work on Wireless Telegraphy is required for the information and guidance of Officers and Men of H.M. Fleet; for this purpose the "Admiralty Handbook of Wireless Telegraphy, 1938" has been prepared at H.M. Signal School.

This book, now in two volumes, supersedes "The Admiralty Handbook of Wireless Telegraphy, 1931."

By Command of Their Lordships,

R.H.A. Carter

.. 1938

1939

PREFATORY NOTE.

THIS Handbook is now divided into two Volumes. Vol. I contains the basic "Magnetism and Electricity Theory" up to and including the "Oscillatory Circuit"; Vol. II contains a treatment of "W/T Theory." The whole is intended to act as a companion book to various sets of purely technical instructions.

It will be found that certain paragraphs, or portions of paragraphs, are marked with an asterisk and usually printed in smaller type, to indicate that they are of a more difficult nature than the rest of the book. Those who are unable to follow the treatment may omit them, without much detriment to the sequence of the argument.

It will be noted that the various sections of Vol. II have been given characteristic reference letters. References in that index are given in terms of paragraph numbers and section letters, those in Vol. I being in terms of paragraphs only.

In order to bring **the unit of capacity** into line with commercial practice, the use of the *jar* as the Service unit of capacity has been discontinued (A.F.O. 1552/37). It is to be considered as obsolescent for a few years, the *farad* and its sub-multiples gradually replacing it as the practical unit of capacity for standard use in the Service.

NOMENCLATURE OF WAVES.

IN this Handbook, the range of frequencies of the æther waves used in wireless communication is sub-divided as follows :—

Below 100 kc/s.	Low Frequencies (L.F).
100 1,500 kc/s.	Medium Frequencies (M/F).
1,500-6,000 kc/s.	Intermediate Frequencies (I/F).
6,000-30,000 kc/s.	High Frequencies (H/F).
Above 30,000 kc/s.	Very High Frequencies (V.H/F).

It has been a common practice in the past to refer to the oscillatory currents produced by these waves in a receiving aerial as H/F currents, and to differentiate them from the currents of audible frequency, produced after detection, by using the term L/F for the latter. It is obvious that this usage conflicts with the nomenclature in the table above; hence in this Handbook the term **radio frequency (R/F)** is applied to all currents directly produced by an incoming signal, and the currents flowing after detection are called **audio frequency (A/F)** currents.

It also frequently happens that an oscillatory current whose frequency falls within the wireless range is generated by the action of a receiver, *e.g.*, in superheterodyne and quench receivers. When describing the action of such receivers it is desirable to distinguish these currents from the R/F currents produced by an incoming signal. The designation of **supersonic frequency (S/F)** currents has therefore been adopted for these currents.

International classification of frequencies would be very desirable. On the basis of a C.I.R. recommendation, promulgated in French, a suitable nomenclature, likely soon to pted internationally, may be given in English as follows :—

Below 30 kc/s.	Very Low Frequencies (V.L/F).
30-300 kc/s.	Low Frequencies (L/F).
300-3,000 kc/s.	Medium Frequencies (M/F).
3-30 Mc/s.	High Frequencies (H/F).
30-300 Mc/s.	Very High Frequencies (V.H/F).
300-3,000 Mc/s.	Decimetre Waves (dc/W).
3,000-30,000 Mc/s.	Centimetre Waves (cm/W).

nce 1st January, 1938, the latter classification has been adopted for standard use in the

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TABLE I.

The Greek Alphabet is here given for reference, as many Greek letters appear in the Text.

Letter.		Name.	English Equivalent.
Small.	Capital.		
α	Α	Alpha	a
β	Β	Beta	b
γ	Γ	Gamma	g
δ	Δ	Delta	d
ε	Ε	Epsilon	ĕ (as in " met ")
ζ	Ζ	Zeta	z
η	Η	Eta	ēē (as in " meet ")
θ	Θ	Theta	th
ι	Ι	Iota	i
κ	Κ	Kappa	k
λ	Λ	Lambda	l
μ	Μ	Mu	m
ν	Ν	Nu	n
ξ	Ξ	Ksi	x
ο	Ο	Omicron	ŏ (as in " olive ")
π	Π	Pi	p
ρ	Ρ	Rho	r
σ	Σ	Sigma	s
τ	Τ	Tau	t
υ	Υ	Upsīlon	u
φ	Φ	Phi	ph
χ	Χ	Chi	ch (as in " school ")
ψ	Ψ	Psi	ps
ω	Ω	Ōmega	o (as in " broke ")

TABLE II.

Symbols for Quantities for Use in Electrical Equations, etc.

Number.	Quantity.	Sign.
1	Length	l
2	Mass	m
3	Time	t
4	Angles	$, \theta, \phi$
5	Work or Energy	W
6	Power	P
7	Efficiency	η
8	Period	T
9	Frequency	f
10	$2\pi \times$ frequency	ω
11	Wavelength	λ
12	Phase displacement	ϕ
13	Temperature, centigrade	t or θ
14	Temperature, absolute	T or Θ
15	Quantity or charge of electricity	Q
16	Current	I
17	Voltage (E.M.F. or P.D.)	E or V
18	Resistance	R
19	Specific Resistance or Resistivity	ρ
20	Conductance	G
21	Specific Conductance or Conductivity	γ
22	Specific Inductive Capacity or Dielectric Constant	K
23	Electrostatic Field Strength	X
24	Electrostatic Displacement or Flux Density	D
25	Electrostatic Flux	ψ
26	Capacity	C
27	Magnetic Pole Strength	m
28	Permeability	μ
29	Magnetic Field Strength	H
30	Magnetic Induction or Flux Density	B
31	Magnetic Flux	Φ
32	Magnetic Reluctance	S
33	Magneto Motive Force	G
34	Self Inductance	L
35	Mutual Inductance	M
36	Reactance	X
37	Impedance	Z
38	Susceptance	B
39	Admittance	Y
40	Base of Naperian logs	e
41	Damping Factor	α
42	Logarithmic Decrement	δ
43	Aerial Capacity	σ
44	Valve mutual conductance	g_m
45	Valve A.C. resistance (impedance)	r_a
46	Valve amplification factor	m
47	Percentage modulation	N
48	Coil amplification factor ($\omega L/R$)	Q
49	Velocity of E.M. Waves	c

TABLE III.

Distinguishing Symbols for Constant and Virtual Values of Quantities.

Number.	Quantity.	Constant Value.	Maximum Value.	Arithmetic Mean Value.	Virtual Value.	Instantaneous Value.
1	Potential Difference ..	V	$\mathcal{V}$	$\overline{V}$	V	v
2	E.M.F.	E	$\mathcal{E}$	$\overline{E}$	E	e
3	Charge	Q	$\mathcal{Q}$	$\overline{Q}$	Q	q
4	Current	I	$\mathcal{I}$	$\overline{I}$	I	i
5	Flux	Φ	Φ_m	$\overline{\Phi}$	Φ	ϕ
6	Magnetic Field	H	$\mathcal{H}$	$\overline{H}$	H	h
7	Electric Field	X	$\mathcal{X}$	$\overline{X}$	X	x

TABLE IV.

Prefixes for Multiples and Submultiples of Quantities.

Number.	Multiple or Submultiple.	Name.	Prefix.
1	10^{6}	Mega-	M
2	10^{3}	Kilo-	k
3	10^{2}	Hekto-	H
4	10^{-2}	Centi-	c
5	10^{-3}	Milli-	m
6	10^{-6}	Micro-	μ
7	10^{-9}	Millimicro-	$m\mu$
8	10^{-12}	Micro-micro	$\mu\mu$

TABLE V.

Signs for Units Employed after Numerical Values.

Number.	Unit.	Abbreviation.
1	Ampere	A
2	Volt	V
3	Ohm	Ω
4	Coulomb	C
5	Joule	J
6	Watt	W
7	Farad	F
8	Henry	H
9	Watt-hour	Wh
10	Volt-Ampere	VA
11	Ampere-hour	Ah
12	Kilowatt	kW
13	Kilo-volt-ampere	kVA
14	Kilowatt-hour	kWh
15	Decibel	db

THE SPARK TRANSMITTER.

1. **The Obsolescence of the Spark Transmitter.**—The radiation of damped wave trains produces interference covering a wide frequency band. Owing to the limited number of available channels of communication in an æther already overcrowded, international regulations have been devised to restrict and, gradually, almost eliminate the use of spark transmission.

At the International Radio Telegraph Convention of Washington, held in 1927, an agreement was reached in the following general terms :—

(*a*) The use of damped wave trains (Type B waves) employing frequencies below 375 kc/s. to be forbidden as from the 1st January, 1930.

(*b*) No new spark transmitting installation to be fitted in a land or fixed station, shore stations being prohibited from using damped waves as from the 1st January, 1935.

(*c*) No new installations for the emission of spark wave trains to be fitted in ships or in aircraft, as from the 1st January, 1930, except when the transmitters working on full power consume less than 300 watts measured at the input of the supply transformer, at audible frequency.

(*d*) Use of spark transmission on all frequencies to be forbidden, as from the 1st January, 1940, except for ship installations, fulfilling the conditions as to power referred to in (*c*).

In the Service, a modern ship installation usually contains :—

(*e*) A spark attachment, for use as a stand-by transmitter in the event of a complete breakdown of the valves or essential components of a valve transmitter. It usually obtains its power supply from the main transformer.

(*f*) An emergency coil, designed for use when power from the ship's mains also fails. It consists of an induction coil and associated simple oscillatory circuits, and it derives its power supply entirely from batteries.

These emergency transmitters are seldom used and are only included in a modern installation in order to provide " a last line of defence."

2. **Methods of Charging up the Condenser.**—In the " oscillatory circuit " chapter of Vol. I we saw that a condenser in series with an inductance and a spark gap, if charged up to a voltage sufficient to break down the insulation of the spark gap, would give a high-frequency oscillatory current in the circuit. This current is damped out at a rate which depends on the losses in the circuit, and by using a suitable type of circuit it may be made to produce an electro-magnetic wave in the æther which carries energy to a distance.

The effective duration of the oscillatory current, and hence of the oscillation in the æther, is very short, and, to render the apparatus effective for communication, it is necessary to have methods by which the condenser may be charged up to discharging voltage at regular intervals. The action mentioned above will then be repeated every time the condenser discharges across the spark gap, and will give a series of wave-trains occurring regularly. It will be seen, when we consider the question of reception, that the most suitable **spark train frequencies** (*i.e.*, the frequencies of the condenser discharges) are those which correspond to the frequencies of easily audible sounds (say, 250 to 1,000 cycles per second), and the various methods of regularly charging and discharging the condenser are designed to work at these frequencies. The methods used, which will be described in this chapter, fall under two headings :—

(1) Direct-current supply methods.

(2) Alternating-current supply methods.

Direct-current methods are only suitable for low-power work, owing to the difficulty of obtaining high-voltage D.C. supplies. They all work on some sort of intermittent make-and-break principle, and will be described in detail under the headings of :—

(*a*) Induction coil.

(*b*) Attracted armature buzzer.

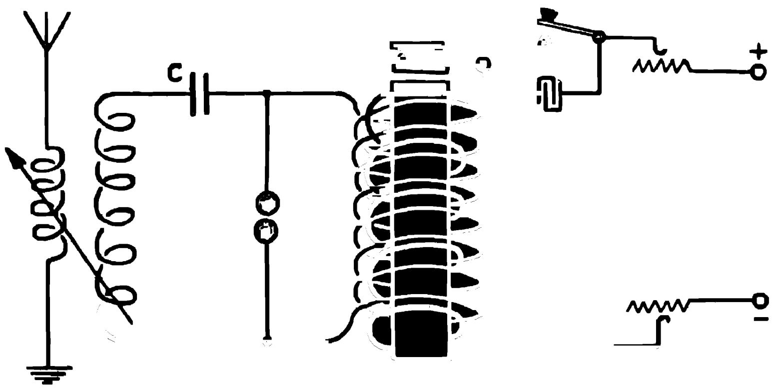
(*c*) Motor-driven buzzer.

The second type, using alternating-current supply, is universally used for large power work. The voltage of the alternator is usually stepped-up by a transformer, so that the method is referred to as :—

(*d*) Alternator and transformer method.

3. **The Induction Coil.**—If it is not convenient to run an alternator, owing to there not being a D.C. supply of high enough voltage or power available, or if it is required to work a small set off a low-voltage D.C. supply or accumulator, the induction coil may be used.

The induction coil may be said to convert a D.C. supply into an intermittent current charging the transmitting condenser at a very much higher voltage than that of the supply.



Induction Coil.

FIG. 1.

The induction coil method is used in the Service in some spark transmitters fitted as attachments to continuous-wave transmitting sets for emergency purposes.

4. **Construction.**—An induction coil consists of a **primary coil** of thick wire wound with a number of turns on an iron core composed of a bundle of soft iron wires.

The primary coil is enclosed in an ebonite tube. Outside this again is the **secondary coil,** which consists of some miles of fine copper wire.

It is built up of a number of flat coils or sections, made by winding the wire between paper discs in a spiral form. The coils are joined in series, inner ends and outer ends of coils being joined together alternately, so that the current flows in the same direction in each coil.

By connecting them in this manner, there is no undue potential strain between the end of one section and the end of the next section to which it is joined.

In series with the primary winding is joined the " Interrupter " or " Make and Break."

This consists of a soft iron armature, secured to the top end of a flat steel spring whose tension can be adjusted by means of an ebonite wheel and adjusting screw.

This armature is close to one end of the iron core, the play between the two being about $\frac{1}{16}$-inch.

The armature carries a small platinum contact. Close to the vibrating armature is a fixed standard carrying an adjustable contact with a platinum tip. These contacts are, normally, held together by the tension of the spring.

Suitable terminals are provided to which the sending key should be joined.

A condenser of large capacity is joined across the key and interrupter contacts ; its functions are explained below.

Two resistances are joined in series with the D.C. supply in order to limit the voltage to a value suitable for working the coil.

5. **Action.**—When the key is pressed, a current flows from the main positive lead, through the key, across the interrupter contacts, through the primary winding, through the resistance and back to negative. The core of the coil is magnetised, and the armature is attracted to it The contacts of the interruptor are therefore suddenly separated, and the current through the primary falls to zero very rapidly.

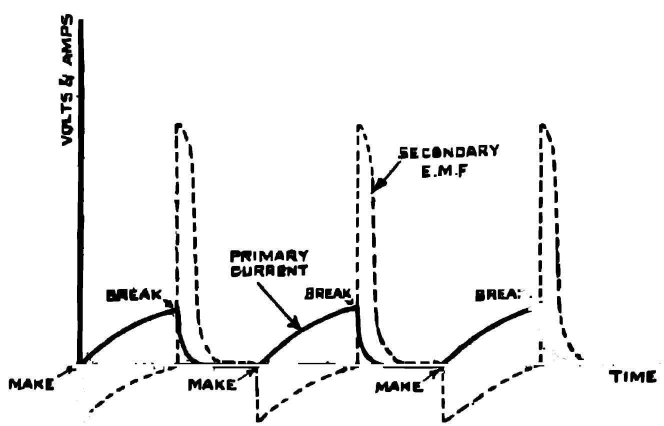
As soon as the primary current has died away, the armature will be released by the core, and will fly back ; and so its contact will make connection again with the fixed contact. The primary current will again start to flow, and the cycle of events be repeated.

6. **Secondary Voltage.**—When the primary circuit is made, as indicated above, the primary current rises comparatively slowly owing to the big self-inductance of the primary coil. There are induced voltages set up in both the primary and secondary windings due to the changing curren and changing flux. In the primary the induced voltage is simply the counter E.M.F. of self-induction the maximum value of which is the value of the impressed voltage (at the beginning of the curren flow, when the current is zero, and the applied voltage is entirely balanced by the induced voltage)

The induced voltage in the secondary while the primary current is growing can, therefore, hav a maximum value which is N times the primary applied voltage, where N is the ratio of the numbe of secondary to primary turns. This secondary voltage is small compared with the voltage induce when the primary current is dying away, as we shall see.

When the contacts of the interrupter are separated, the primary current falls to zero. Agai there are voltages induced in both the primary and secondary circuits due to changing curren and changing flux, but the presence of the large-capacity condenser results in the rate of change c current and flux being much greater than during the period when the current is growing. Th secondary E.M.F., being given by the product of the number of turns and **the rate of change c flux,** is therefore much greater than during the growth of current.

7. **Condenser Action.**—(1) The first function of the condenser across the interrupter is t minimise the sparking which tends to occur at the interrupter contacts. When the primary circu



Primary Current and Secondary Voltage in Induction Coil.

FIG. 2.

s interrupted, the counter E.M.F. of self-induction due to the current decreasing would tend to keep the current flowing across the small gap that forms. Instead of this happening, the counter E.M.F. charges up the shunting condenser which had previously been short-circuited by the interrupter, and which, being of large capacity, does not rise to a high potential. The potential across he gap is therefore limited, and sparking is avoided. The break between the contacts is cleaner and a higher E.M.F. is induced in the secondary.

(2) As the condenser discharges, its discharge current opposes the primary current and helps t to die away. The circuit in which the primary current decays is really a highly damped oscillatory circuit and if the resistance is not too large, a damped oscillation will be set up. In any case, the rate of decay of current and flux is increased.

For these two reasons, the E.M.F. induced in the secondary coil during the period of decay of he primary current is much greater than that induced during the period of growth of the primary current, the determining factor being the rate of change of current.

In both cases, of course, the secondary voltages depend on the amplitude of the primary current and its attendant flux, and the number of turns in the secondary.

A figure is appended, showing the variation of primary current and induced E.M.F. in the secondary circuit.

8. **The Oscillatory Circuit.**—Let us connect the secondary of an induction coil to an oscillatory circuit, consisting of a condenser, an inductance, and a spark gap. When the interrupter breaks the primary circuit, the very high induced E.M.F. in the secondary charges the condenser to such a high voltage that a spark takes place between the balls.

The high resistance of the gap having now been bridged by a spark, the condenser discharges hrough the inductance and the spark gap, and a high-frequency oscillatory current is set up in the circuit, and energy is radiated.

This action takes place at each break of the interrupter.

The frequency at which wave-trains are radiated is determined by the mechanical constants of the system—the tension of the spring and the position of the armature, etc

9. **Limitations of an Induction Coil.**—Although an induction coil may produce a momentary voltage of as much as 150,000 volts, yet it is not a very effective means of charging up a condenser for the reason that the **duration** of the high secondary voltage is very short.

The charging circuit consists of a condenser C being charged through a resistance R_s, the resistance of the secondary winding.

It was seen in Vol I that, in such a case, the time taken to charge up the condenser is proporional to CR_s, the time constant of the circuit

Since the duration of the high secondary voltage is short, CR_s should be small to enable the condenser to be charged to a high potential. C is fixed by H/F oscillatory conditions, and therefore R_s should be small The secondary winding must, however, have a large number of turns of fine wire, and so R_s in practice is large. Hence an induction coil is only suitable for charging up a comparatively small condenser ; the energy isolated in the oscillatory circuit for each wave-train s small and the amount of energy radiated is correspondingly small.

This matter may be described somewhat more precisely as follows :—

The energy stored in the field of the primary winding is given by—

$$W = \tfrac{1}{2} LI^2 \text{........where I is the battery current.}$$

At the " break," the field collapses and the stored energy is transferred to the secondary winding within the " period " of the hammer.

It will be assumed that the effect of this is to produce a large charging current " i " that may e regarded as constant for the short time " t," which is governed by the rapidity of the break.

Some of the transferred energy will be dissipated in ohmic losses, and some will go to the condenser, hence

$$W = \frac{1}{2} CV^2 + i^2 R_r t.$$

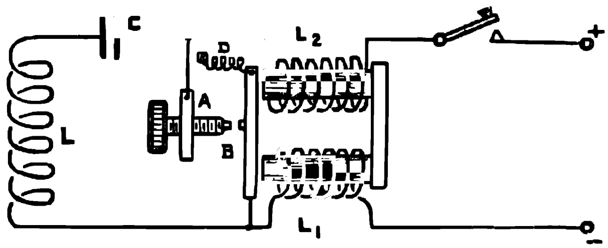
But $i\,t = Q$, the charge in the condenser, and $Q = CV$.

Hence, $$W = \frac{1}{2} CV^2 + \frac{C^2V^2R_r}{t}$$

$$\therefore \text{ Energy in condenser} = \frac{1}{2} CV^2 = \left(\frac{W}{1 + \frac{2CR_r}{t}}\right) \quad \text{(1)}$$

Expressing t in terms of the time constant; if it were $2CR_r$, the energy stored in the condenser would be only half of that available in the primary winding.

10. **The Attracted Armature Buzzer.**—Another method of energising an oscillatory circuit is by means of a buzzer. This is illustrated in Fig. 3.



Typical Buzzer.

FIG. 3.

Here we have a fixed contact A, and a contact B carried on an armature which is free to vibrate. A and B are, normally, held together by a spiral spring D.

Joined to B and A are two coils of high inductance, L_1 and L_2.

This arrangement constitutes the buzzer. A D.C. source of supply is joined to L_1 and L_2, and is made and broken by a signalling key.

Across the make and break AB is joined a circuit, LC.

11. **Action.**—(*a*) When the key is pressed, current flows through L_2, across the contacts AB, and back through L_1. The cores of L_1 and L_2 are magnetised and attract the armature B away from A.

(*b*) When the D.C. circuit is suddenly broken at AB, the current cannot suddenly cease owing to the high inductance of the coils L_1 and L_2, so that the condenser C is charged up by the inductive kick from L_1 and L_2, the right-hand plate being charged positively and the left-hand one negatively.

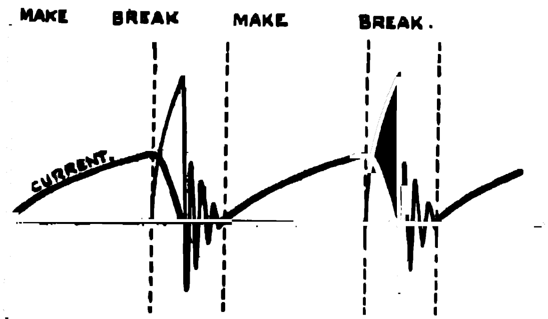


FIG. 4.

The energy that was stored in the magnetic fields round L_1 and L_2 is transferred to C when the current in L_1 and L_2 has been reduced to zero.

(*c*) The cores of L_1 and L_2 being demagnetised, the moving contact flies back towards the fixed contact.

When the two contacts are nearly together again, the P.D. between the condenser plates causes a spark to pass across the small gap so formed, and the discharge of the condenser C sets up a high frequency oscillation in the circuit formed by the condenser C, the inductance L and the small spark gap between A and B.

(*d*) When A and B touch once more, a direct current will flow through L_1 and L_2, and the operation will continue.

This action may be summarised as in Fig. 4.

The thick line illustrates the slow rise of the direct current through L_1 and L_2 at the moment of "make," because of their self-inductance; and the sudden fall of the direct current at the moment of "break," because of the P.D. across the condenser C at this moment. This is tending to send a current in the opposite direction to that flowing in the inductances and so the latter current decays more quickly than it rose. The condenser in the oscillatory circuit performs, from this point of view, the same function as the shunting condenser used with the induction coil.

The thin line shows the voltage applied to the condenser C at the moment of "break" and the high frequency oscillation set up in the circuit LC.

It is important to arrange the mechanical constants of the circuit correctly, so that, when the two contacts are nearly together and a spark is due to take place, the condenser will be charged up to its maximum voltage.

12. The above method of energising an oscillating circuit has been used extensively :—

(*a*) To energise a transmitting oscillator, when only a comparatively short range is required.

(*b*) To energise receiving circuits for testing or tuning purposes.

(*c*) For laboratory work.

An advantage of this method of energising a transmitting oscillator is that, with reasonably loose couplings, only one wave is emitted from the aerial, and not two waves (due to interaction of primary and aerial circuits) as previously described.

The spark in the primary oscillatory circuit is extinguished in an exceedingly short time because the large mass of metal in its vicinity is very efficient in conducting away the heat which normally sustains the ionisation in the gap. Thus the primary is put on open circuit and no energy is transferred back to it from the aerial circuit. The high-frequency oscillation continues in the aerial circuit alone at the natural frequency of the latter. This type of action is known as "**quenching**," and will be considered later with respect to a special type of spark gap which effects it very efficiently.

13. **Another Type of Buzzer.**—The circuit is shown in Fig. 5. When the switch is made there is a closed circuit through the oscillatory inductance, buzzer coil and buzzer contacts. The buzzer coil is energised and separates the contacts.

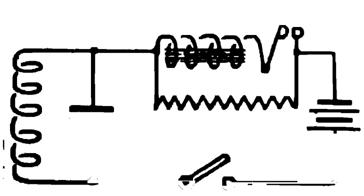


FIG. 5.

At the "break" of the current, the size of the E.M.F. set up across the buzzer coil is limited in value by means of the parallel resistance of 100 ohms. The effect is to produce a small but rapidly quenched spark between the buzzer contacts. The hammer end of the buzzer coil tends to remain at positive battery potential (Lenz's law), and the other end tends suddenly to fall below the potential of the negative end of the battery, thereby causing the condenser to become charged and a damped radio-frequency oscillation takes place in the LC circuit.

It should be noted that the P.D. limiting resistance is short-circuited through the buzzer coil when the contacts are together and, therefore, does not reduce the exciting current from the battery.

14. **The Motor-driven Buzzer.**—A **motor buzzer** is a more efficient method of energising an oscillating circuit than the attracted armature type of buzzer.

It may be illustrated diagrammatically as follows :—

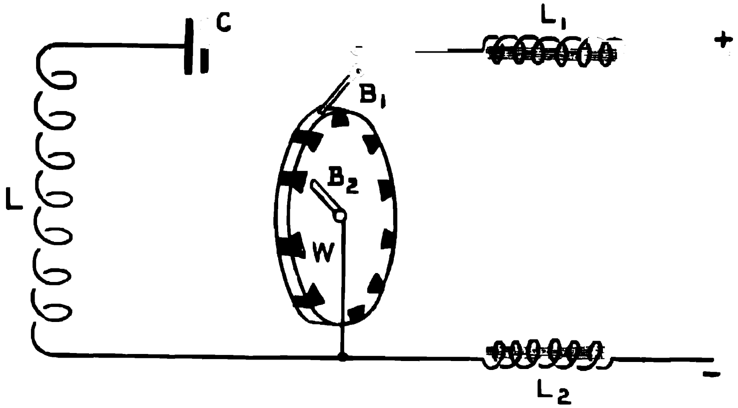


FIG. 6.

W represents a wheel driven round at a high speed by a small motor. In its edge are set a number of insulating segments of mica.

Bearing on its edge and on its side are two brushes, B_1 B_2, across which is joined the oscillating circuit LC.

Joined to the two brushes are two coils of large inductance L_1 and L_2.

A D.C. source of supply, interrupted by a hand key, is connected to the two inductance coils.

15. **Action.**—The action of this type of buzzer is very much the same as that of the attracted armature type previously described.

(*a*) When brush B_1 is bearing on a **conducting** segment of the wheel a steady current will flow through L_1 and back through L_2.

(*b*) When the **insulating** segment on the wheel comes under brush B_1, connection between the brushes is broken. The counter E.M.Fs. of L_1 and L_2 (due to the magnetic fields set up round them by the current) will charge up condenser C.

(*c*) When the gap between brush B_1 and the next edge of the conducting segment is small enough, the condenser C will discharge in a high frequency oscillation, a small spark gap being formed across a portion of the surface of the insulating segment.

This sparking should occur on the under side of the brush. If it occurs above the brush it indicates that the wheel is dirty or that excessive power is being used.

Fig. 4 will serve to illustrate also the action of the motor buzzer. The period marked " make " is that during which the brush is bearing on a conducting segment, and the period marked " break " that during which the brush bears on an insulating segment.

This design of buzzer, like the previous type, produces a **Quenched Wave.**

With this type of buzzer, as with the attracted armature buzzer, it is important to arrange the circuit so that the current falls to zero, and the condenser C is therefore charged up to its maximum voltage, when the gap between the brush and the next conducting segment is small enough for a spark to take place.

During the period of break the circuit L_1, L, C, L_2 is effectively an oscillatory circuit with a steady source of supply ; the current will fall to zero and the condenser charge up, in a period which is about a quarter of the **natural** period of oscillation of **this** circuit. L_1 and L_2 should be so arranged, therefore, that the duration of " break " should be just less than this quarter-period The condenser will then be at its maximum voltage when the spark is due to take place.

Power may be reduced by inserting series resistances in the supply circuit.

16. The brushes should bear as firmly as possible on the surface of the wheel, and the standards on which they are mounted should have no play in them.

Excessive sparking at the brush, loose brush holders, or a pitted or dirty wheel may cause a bad note and loss of range.

The condenser is charged and discharged every time the brush bears on an insulated segment. Therefore, the number of times per second the condenser is charged, is equal to the speed of the wheel, in revolutions per second, multiplied by the number of insulating segments.

17. **The Alternator and Transformer Method.**—The methods hitherto described for charging up a transmitting condenser are only used under the special conditions referred to.

The almost universal practice for energising spark oscillatory circuits of $\frac{1}{2}$ kW. sets and upwards is to use an alternator or rotary converter and transformer.

It is essential to charge the condenser up to a very high voltage, for otherwise an excessively big condenser would be needed to store up a reasonable quantity of energy (Energy $= \frac{1}{2}$ CV^2), which would mean that only large LC values could be obtained in the oscillating circuit. We might produce this voltage directly from some **high voltage alternator**, but such machines are very difficult and expensive to construct, and are not at all efficient ; also, the whole circuit would have to be very carefully enclosed to prevent fatal shocks being taken off it.

Fortunately, it is a very easy matter to increase the voltage of an alternating current by utilising a step-up transformer.

Our simplest arrangement is then as shown in Fig. 7, namely, a low voltage alternator (or rotary converter) of a power suitable for the set we are using, delivering its alternating current at a frequency depending on the number of its poles and the speed at which it is run.

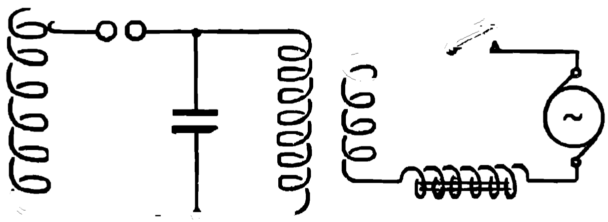


FIG. 7.

The circuit as shown may be divided into—

(*a*) Low Tension circuit, and High Tension circuit.

(*b*) Charging or Low Frequency circuit, and oscillatory or High Frequency circuit.

(*a*) **Low-tension Circuit.**—This includes everything to the right of the transformer in Fig. 7. From the terminals of the alternator, current will flow through the primary of the transformer and across a signalling key of some sort. The impedance coil will be described later.

High-tension Circuit.—This includes the secondary of the transformer and everything to the left of it in Fig. 7. The secondary terminals are connected to the oscillatory circuit, which consists of a condenser, spark gap, and inductance. The current flowing from the secondary terminals of the transformer will be a **small** one at a **high voltage,** and at the same frequency as that of the alternator.

(*b*) **Charging Circuit.**—In Fig. 7 this means the condenser and everything to the right of it.

Oscillatory or H/F Circuit.—This means the condenser, spark gap and inductance.

Note.—As regards the arrangement of the oscillatory circuit, it is a matter of indifference whether we take the transformer leads to each side of the condenser, or to each side of the spark gap, as in Fig. 8.

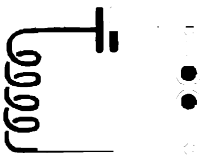


FIG. 8.

In the latter case the charging current flows through the transmitting inductance. Both arrangements may be found in practice.

18. **The Charging Circuit.**—It has been customary to consider the charging circuit from the point of view of resonance, and to show theoretically that the condenser will be charged up to the maximum voltage if the charging circuit is made resonant to the frequency of the alternator. On this assumption the impedance coil was regarded as a variable inductance which tuned the charging circuit, the latter being considered as a single circuit in accordance with the theory by which inductance or capacity on the secondary side of a transformer can be given an equivalent value and transferred to the primary side (*see* Vol. I). In sets still in use in the Service, however, the LC value of this equivalent single circuit, even with the minimum setting of the impedance coil, is always greater than the CL value which would correspond to the frequency of the supply, so that the theory of resonance of the charging circuit is not borne out in practice. It will be seen that the function of the impedance coil in the practical case is really to act as a regulator of the voltage impressed on the primary of the transformer, and in addition to assist in preventing what is known as "arcing."

19. **Condenser Voltage.**—Let us assume that the circuit is made at the beginning of a cycle of alternator voltage. A current will start to flow in the charging circuit and the voltage across the condenser will start to build up. From the theory at the end of the "Oscillatory Circuit Chapter" (Vol. I) the conditions at first will be "transient"; in other words, the graphs of current and voltage to begin with will not be sinusoidal, but will be the resultant of a sine wave (representing the final conditions of forced oscillations), and a damped wave at the natural frequency of the circuit (representing free oscillations).

The complete wave form will be of the type shown in Fig. 9.

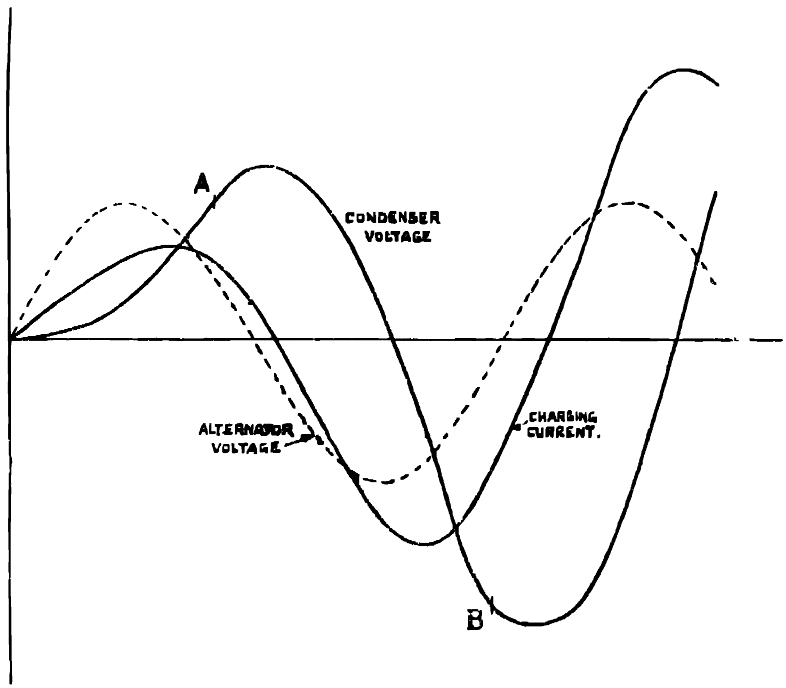


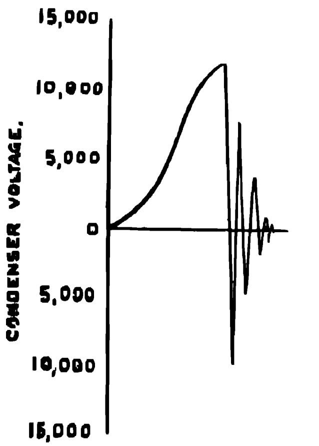
FIG. 9.

When the transient stage is over, the curve representing current will be a sine wave, lagging on the alternator voltage by a definite phase angle, while the voltage across the condenser will also be sinusoidal in form and lagging by 90° on the current.

These final conditions are not, however, of importance, because the spark gap is adjusted so that it breaks down at some voltage less than the final maximum value, for example, at point A in the figure above; under these circumstances, the transient conditions recur again after the spark gap breaks down, and the initial current and voltage variations are repeated.

20. **Spark Train Frequency.**—Let us consider in detail what happens when a spark occurs. We shall take first the simple case of a fixed spark gap. By this is meant a gap in which the plugs are fixed relatively to each other while sparking is taking place, but whose distance apart can be varied before the transmitting key is pressed. Let the spark gap be set at such a distance that the voltage across the condenser, and hence across the gap, corresponding to that at point A (Fig. 9), breaks down the insulation of the gap.

An oscillatory action takes place in the high-frequency circuit, the spark gap being now equivalent to a resistance. Energy is transferred by means of the mutual coupling to the aerial circuit, and a certain proportion is radiated into space in the form of electro-magnetic waves. The time during which the oscillatory action occurs is very small indeed compared with that of one cycle at the frequency of the alternator, so that the curve representing voltage across the condenser during the charging period and the consequent oscillatory discharge will be somewhat as shown.



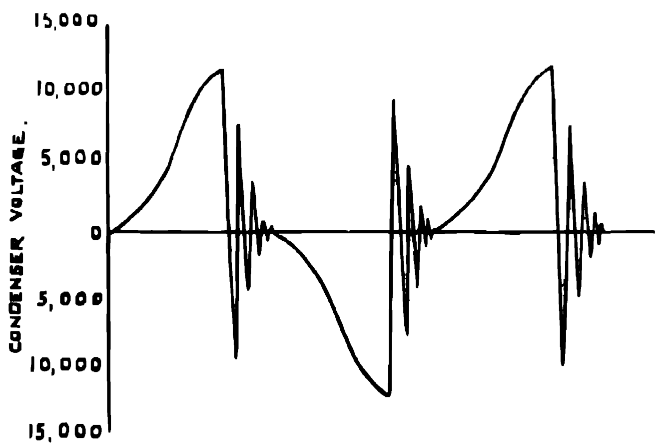
Condenser Voltage under Discharge Conditions.

FIG. 10.

The voltage across the condenser when the discharge is just completed is, of course, zero. If the break-down voltage is so arranged that, at the end of the ensuing discharge, a cycle of alternator voltage is just commencing again, exactly the same sequence of events will be repeated.

The case we have taken, of spark gap breakdown before the condenser voltage had reached its first maximum value, and the completion of the oscillatory action when the alternator voltage is again zero, as at the beginning, will obviously give **one spark per half-cycle** of the alternator voltage.

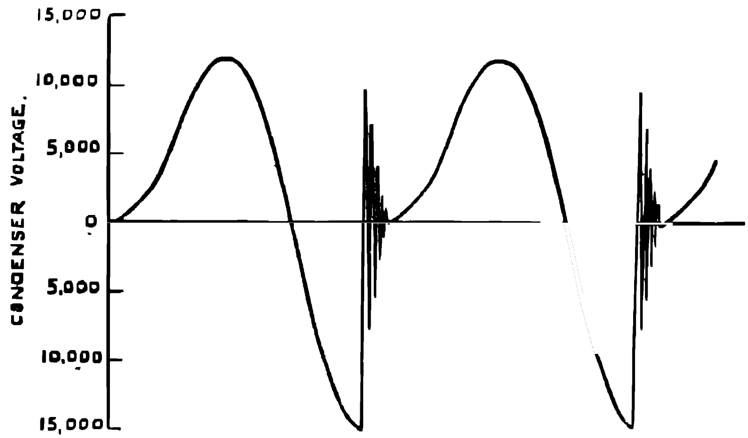
The sequence of events during succeeding half-cycles may be illustrated by the following graph of condenser voltage. Spark trains have been drawn to end where the next half-cycle of alternator voltage is beginning, although this will not in general be the case.



One Spark per Half-cycle.

FIG. 11.

If the spark gap were set at such a distance that the voltage at point B (Fig. 9) was just sufficient to break down its insulation (this voltage being greater than the first peak value), we should get **one spark per cycle** instead of one spark per half-cycle, and the curve representing condenser voltage would be of the form shown in Fig. 12.



One Spark per Cycle.

FIG. 12.

It is important clearly to understand the difference between the two frequencies involved in a spark transmitter, and to appreciate the shortness of time during which energy is being radiated into space in comparison with the case of (say) a C.W. transmitter.

Assume a spark train frequency of one spark per cycle, of numerical value 500 times per second. Assume, also, that the LC value of the oscillatory circuit corresponds to a frequency of 500 kc/s. The condenser receives a charge from the alternator and transformer once in every 1/500 second. If we assume that there are 100 complete oscillations in the oscillatory discharge of the condenser the time taken before the condenser is fully discharged is 100/500,000 second, since each cycle takes 1/500,000 second. The total time of discharge is thus 1/5,000 second, and is therefore only a tenth of the time (1/500 second) between condenser charges.

Thus, for nine-tenths of the time that the key is pressed, no energy is being sent out into space as æther waves, which is one reason for the comparative inefficiency of spark transmission. Fig. 13 (*b*) attempts to show this to scale.

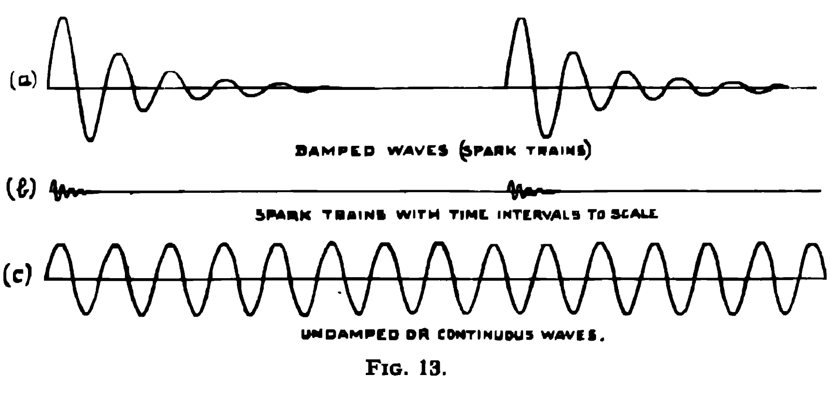


FIG. 13.

With the type of gap which was in common use in the Service, the asynchronous rotary gap, these curves, which give a regular recurrence of events, have to be modified. This type of gap will be treated later.

21. **Arcing and the Impedance Coil.**—When the insulation of the spark gap is broken down, and the high frequency oscillatory current is flowing across the gap, the intense heat generated by the spark at the moment of discharge is sufficient to volatilise some of the metal of which the spark gap electrodes are made. This volatilised metal forms a conductive bridge from one electrode to the other. So long as a certain minimum current is passing across the gap, this conductive bridge is maintained. Now, when the spark gap becomes a conductor, the current from the alternator, which has hitherto been charging up the condenser, will tend to flow across the gap, maintaining its conductive property and preventing the condenser from being charged up again. It is therefore necessary to take every possible precaution for restoring the insulating properties of the spark gap as soon as the high-frequency discharge is completed.

Apart from mechanical devices for prevention of arcing, such as an air blast, which will be discussed later in dealing with the different types of gap, the inductance of the charging circuit, coupled with that of the impedance coil, tends to prevent charging current flowing across the gap. When the gap becomes conductive, the condenser is short-circuited and the charging circuit may be considered to consist of inductance and resistance only. The high inductive reactance cuts down the current to such an extent that it is insufficient to maintain the conductivity of the gap; the latter being restored, the condenser comes into the charging circuit again and the current charges up the condenser once more.

The **Impedance Coil** is therefore useful as an added inductance in the charging circuit for the prevention of arcing.

A further use of the impedance coil is in controlling the power taken from the alternator. The larger the inductance inserted in the circuit by the impedance coil, the further from resonance are the conditions in the charging circuit; the more is the charging current cut down, and the less is the voltage built up across the condenser. The impedance coil may therefore be looked on as a voltage regulator, controlling the voltage in the condenser and hence the power radiated.

22. **Power taken in Charging a Condenser.**—When a condenser, C farads, is charged up to a discharging voltage of V volts, the amount of energy stored up in it is $\frac{1}{2}CV^2$ joules.

This energy is dissipated in damping losses in the primary and aerial circuits while the condenser is discharging in a high-frequency oscillation across the spark gap.

If we charge up the condenser to a voltage of V volts N times per second (*i.e.*, if N is our spark train frequency), then we shall be expending energy at the rate of $\frac{1}{2}NCV^2$ joules per second. But a joule per second is a watt, the unit of power.

Hence the power required to charge a condenser of C farads to a voltage of V volts N times per second is $\frac{1}{2}NCV^2$ watts.

Expressing power in kilowatts and capacity in jars, the equivalent formula is :—

$$\text{Power in kW.} = \frac{1}{2} \times N \times \frac{C}{9 \times 10^8} \times V^2 \times \frac{1}{1000}$$

$$= \frac{NCV^2}{1{\cdot}8 \times 10^{12}}$$

Example.

Find the power required to charge a condenser of 200 jars to a voltage of 15,000 volts, at a frequency of 300 cycles per second, with a spark train frequency of one spark train per cycle.

$$\text{kW.} = \frac{200 \times 15{,}000^2 \times 300}{1{\cdot}8 \times 10^{13}} = 7{\cdot}5.$$

From the above formula, we can deduce that the—

(*a*) greater the size of the condenser charged,
(*b*) higher the sparking voltage,
(*c*) higher the spark train frequency,

the more is the power required to charge the condenser and the more is the power passed to the aerial and radiated away every second, the preponderant factor being V.

In Service sets the condenser in the oscillatory circuit is composed of separate elements, which may be joined in series or in parallel. The parallel setting is used when a high LC value in the oscillatory circuit is necessary, *i.e.*, on long wavelengths or low wave frequencies. In order to keep as much equivalence as possible in the power radiated on different wave frequencies, the voltage to which the condenser is charged up should be smaller the greater the value of C. As the transformer secondary is in two halves, which may also be joined in series or in parallel, the voltage across the condenser can be reduced by using the parallel connection, or increased by using the series connection. This explains the rule employed in these sets :—

Condensers in parallel, transformer secondaries in parallel.

Condensers in series, transformer secondaries in series.

23. **Design of Spark Gaps.**—Charging circuits are very similar in their general arrangement, differing in detail only according to the power of the set and the range of waves required.

It is chiefly in **design of spark gaps** that sets differ.

The objects aimed at may be classified as follows :—

(*a*) **To Prevent Arcing.**—The effect of arcing has been already discussed.

The object is so to cool the conductive bridge that it is extinguished as soon as the oscillatory discharge is over.

Apart from the effect of the inductance in the charging circuit, and the additional inductance of the impedance coil, various methods of preventing arcing are included in the designs of the different types of spark gap.

(*b*) **To Effect Quenching.**—Quenching is effected by devising a very much more rapid method of cooling the spark gap and restoring its insulating properties than any method used for prevention of arcing. It effects this cooling so quickly that not only is arcing prevented, but the interaction and exchange of energy between the primary and aerial circuits is stopped as soon as the energy has been transferred for the first time to the aerial circuit. The theory of quenching will be considered in paras. 27 and 28, which deal with the special type of gap designed to effect this action.

(*c*) **To Produce a High Spark Train Frequency**.—The pitch of the note heard in the telephones at the receiving station depends on the spark train frequency of the transmitting station. If a fairly high frequency alternator is used, the conditions of one spark per cycle will give a high note.

If, however, the alternator or rotary converter is a low frequency one, we may increase the spark train frequency by using such a short spark gap as to give one spark per half-cycle or even more.

The higher the spark train frequency the more difficult it is to prevent arcing, because the electrodes become hotter. With one type of gap, the asynchronous rotary gap, the spark train frequency is independent of the alternator frequency, and can be adjusted to any value desired.

The different types of gap employed are :—

(*a*) The fixed gap.
(*b*) The synchronous rotary gap.
(*c*) The asynchronous rotary gap.
(*d*) The quenched gap.

These will now be considered in detail. The last two are the types used in Service sets.

24. (*a*) **The Fixed Spark Gap.**—By this is meant a gap in which the plugs are fixed relatively to each other while sparking is taking place, but can be set to any required distance apart before the transmitting key is pressed.

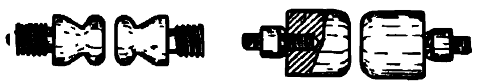
Various shapes are illustrated in Fig. 14.

Such gaps are quite suitable for low power, or for low sparking frequencies, and are simple, requiring little attention beyond a periodical cleaning.

Two waves are emitted from the aerial circuit, the amount by which their frequencies differ depending on the coupling between the primary and aerial circuits.

The gap electrodes should be constructed of non-arcing material.

During the time the gap is conductive its resistance should be low, therefore the distance apart of the electrodes must not be too great. If, however, the distance apart is too short, the tendency towards arcing will be greater. The sparking surfaces should be parallel and clean, otherwise the spark will jump across at the same point, causing the surfaces to be burnt away.



(*a*) (*b*)

FIG. 14.

For higher powers or higher spark train frequencies a **Blower** is often provided to prevent arcing troubles. This is an arrangement for forcing a stream of cooling air between the electrodes. In relatively low power work, a comparatively low pressure stream provided by a fan will usually be found to be sufficient ; at higher powers, it has sometimes been the practice to drill a cylindrical hole through one of the electrodes, and blow air through it at a pressure of several atmospheres in order to effect very rapid quenching.

25. (*b*) **The Synchronous Rotary Gap or Discharger.**—A **Synchronous Rotary Gap** denotes a form of design in which a metal wheel carrying a number of studs or spokes projecting from its edge rotates between two fixed electrodes, the rotating wheel being rigidly fastened to the alternator shaft. When the studs are opposite the fixed electrodes, the spark jumps from one electrode to the wheel stud, through the wheel, and back through the second gap to the other electrode.

As the speed of the wheel and the alternating frequency both depend on the speed of revolution of the motor driving the alternator, the number of times per second at which the condenser voltage reaches a peak value and the number of opportunities it has of discharging can be made equal, and the position of the studs arranged so that these conditions occur simultaneously.

As the number of cycles per revolution of the alternator is equal to the number of pairs of its poles, and as in one revolution of the spark wheel the number of opportunities of sparking is equal to the number of studs on the wheel, it follows that to get **one spark train per half-cycle** the wheel should have as many studs as the alternator has poles.

The position of the fixed studs can be altered relatively to the moving studs by mounting them on a rocker which can be moved in one direction or the other. The condenser can thus attain its peak value of voltage when the studs are exactly opposite each other, and the gap separation is adjusted for the breakdown to occur just before this happens.

Arcing is prevented by the increasing separation of the electrodes as the wave train takes place, and also by the fanning and cooling caused by their rapid motion.

In larger sets, air blasts are also used to cool the electrodes and prevent arcing.

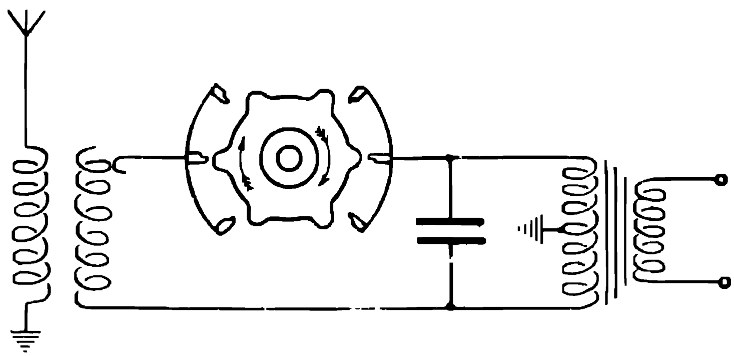
It should be noted that the motor buzzer wheel, as previously described, is only a special variety of a synchronous gap. As the supply is a direct-current one, the spark train frequency is entirely dependent on the speed of revolution.

26. (*c*) **The Asynchronous Rotary Gap.**—This is the type of gap used in some spark sets in the Service.

The principle of the rotating wheel with studs, and the fixed electrodes between which it rotates, is exactly the same as for the synchronous gap, the essential difference in the action being that the **speed of rotation of the wheel is entirely independent of the speed of the alternator.** The

wheel is driven by a separate motor. The word " asynchronous," which means " out of time with," refers to this independence of speed of rotation.

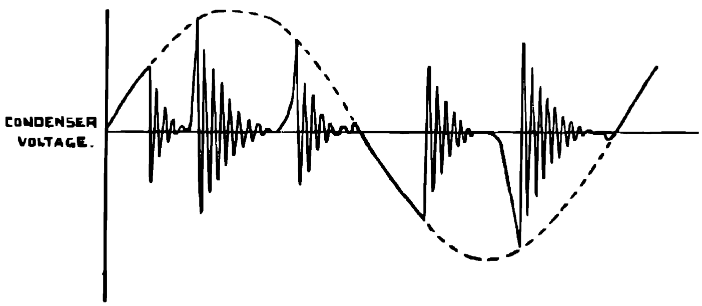
The gap between the fixed and moving electrodes is adjusted to be as small as possible. A spark will occur whenever the gap between a fixed electrode and the point approaching it is small enough for the condenser voltage to force a spark across it. The great advantage of the asynchronous gap is that it is possible to produce a high spark train frequency from a low frequency supply. This being the case, it is obvious that there may be several sparks during one cycle of alternator supply,



Asynchronous Rotary Gap.

FIG. 15.

and consequently these occur at different points in the cycle. The conditions are not exactly repeated each time as in the case of the synchronous spark, because the charging current from the alternator is charging up the condenser during different parts of its own cycle of variation, and hence neither the voltage to which the condenser is charged nor the voltage of breakdown is constant. Since the gap is adjusted to be as small as possible, there will be a discharge each time the studs come opposite each other unless the condenser voltage is very small. Sometimes, however, this condition holds and a spark is missed.



Condenser Voltage with Asynchronous Rotary Gap.

FIG. 16.

Not only is it possible to miss a spark altogether, but the interval between sparks is not absolutely constant. If the charging current after one breakdown is such that the condenser voltage before the next discharge is greater than usual, the spark will take place over a longer gap, *i.e.*, when the studs are a little further apart than usual. If the condenser voltage is less than its average value, the spark will take place a little later than usual.

In addition, the energy stored in the condenser and the proportion radiated in the separate wave-trains is variable.

The disadvantages of this type of gap—the possibility of a spark being missed, irregular time-intervals between sparks, and irregular amplitudes of wave-trains—result in the note heard at a receiving station being impure.

The great advantage is the high spark train frequency which can be got from a low frequency supply, giving more energy radiation and a high and easily readable note at the receiver.

Arcing is prevented, as in the synchronous gap, by the mechanical separation of the electrodes and the draught of air. If necessary, an air blast may be fitted in addition.

A figure is given which shows the way in which the condenser voltage may be considered to vary during one cycle of low-frequency alternator supply, several spark discharges taking place during this period.

27. (*d*) **The Quenched Gap—Theory of Quenching.**—The **Quenched Gap** is designed to put out the spark much more quickly than any of the methods used for prevention of arcing. It was seen in the Chapter on the " Oscillatory Circuit " (Vol. I) that when the primary oscillation has transferred its energy to the aerial, the aerial re-transfers it to the primary, and the energy is transferred and re-transferred backwards and forwards between the two circuits until it has been entirely expended in various damping losses, as illustrated thus :—

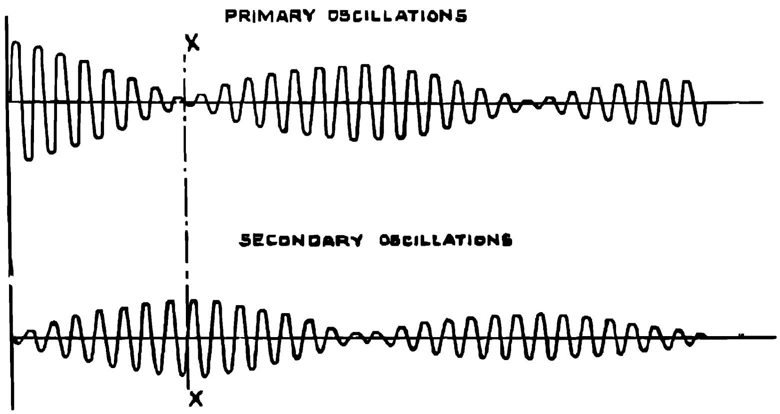


Diagram of Primary and Aerial Currents in Ordinary Coupled Circuit.

FIG. 17.

This transfer and re-transfer of energy produces two harmful results :—

(*a*) The heavy damping in the spark gap is wasting the energy of the high-frequency oscillation during the whole time of transmission.

(*b*) Two waves are emitted which interfere with ships working on other waves of different frequency.

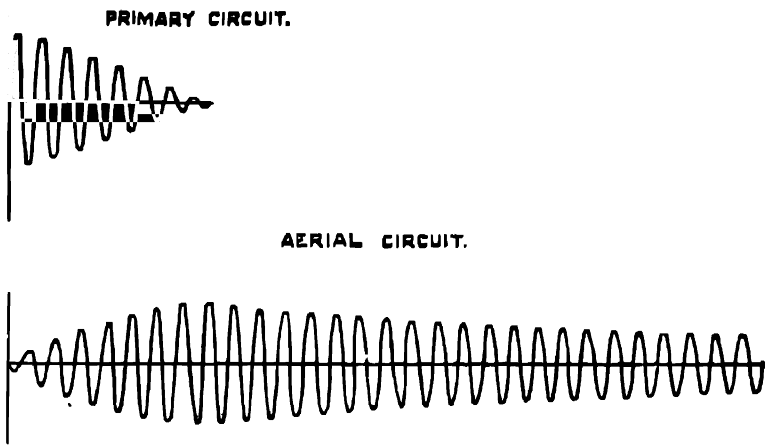
If, in some way, we can manage to put the spark out at the moment X in Fig. 17 (when the primary energy is at a minimum), then the conditions are entirely changed.

As soon as the whole of the energy has been transferred to the aerial the conductivity of the gap is destroyed and the energy which is now in the aerial cannot return to the primary, and will therefore continue to oscillate in the aerial circuit until the whole of it has been radiated in the form of electro-magnetic waves or lost unavoidably in resistance in the aerial circuit.

The coupling between the circuits can be increased, and more energy transferred from the closed to the open circuit. After the spark is " quenched," the aerial circuit oscillates at its own natural frequency, and it is only for a comparatively small period during the transfer of energy that two-wave frequencies are being radiated. These are both considerably different from the natural

frequency of the single circuit, especially with the tight coupling employed, and cause interference at the beginning of the wave-train. The tight coupling causes the aerial current to build up very rapidly, causing a shock effect by setting up free oscillations in neighbouring aerials.

A diagram of the current oscillations in the two circuits is as shown in Fig. 18.



Primary and Aerial Currents in Quenched Gap Circuit.

FIG. 18.

The upper diagram represents the oscillations in the primary circuit until the spark is "quenched" at the moment X (*see* Fig. 17), and the lower diagram those in the aerial circuit. The advantages and disadvantages of quenching may be summarised as follows :—

Advantages.

(*a*) Only one wave frequency radiated for the greater part of the wave-train.

(*b*) No tendency to arc. The time involved in quenching is probably that corresponding to 3 or 4 cycles of high-frequency oscillation, which is very small compared with the time in which anti-arcing devices become efficient. One spark per half-cycle can therefore be obtained quite easily with the quenched gap.

(*c*) The comparatively high resistance of the gap is in circuit for a very short time. As the unwanted power loss in the aerial circuit can be made considerably less than that in the primary circuit, a greater proportion of the energy input is converted to radiation energy.

(*d*) Tight coupling (up to 20 per cent.) can be used, and therefore rapid transference of energy is ensured. The tighter the coupling the less is the time of transference and the less are the losses in the primary circuit.

Disadvantages.

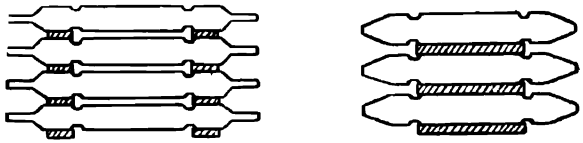
(*a*) The tight coupling causes the aerial oscillation to start with a very rapidly increasing amplitude which shocks neighbouring aerials into oscillation.

(*b*) During the period of interaction two wave frequencies are radiated, each differing from the frequency of free oscillation of the aerial.

Quenching is achieved by the use of the special design of gap called the **Quenched Gap,** which will now be considered in detail. The rapid mechanical rotation of the motor buzzer is sometimes considered to give a quenching action.

28. (*a*) **The Quenched Gap—Construction.**—The Quenched Gap, as now described, is used in spark attachments to main valve transmitting sets in the Service. Quenching is simply a matter of cooling the spark path rapidly enough.

A method of accomplishing this is to make use of the property metals have of conducting heat. In the quenched gap, the spark gap is broken up into a number of very short gaps in series with each other. The electrodes are made of copper, plated with silver, these metals being good heat conductors. In addition, large cooling or radiating fins are provided, so that there is a large mass of metal surrounding the gap and plenty of metal surface exposed to the surrounding air. In addition, it may be necessary in some cases to employ an air blast to cool the fins sufficiently. The gaps are so short that no part of the air dielectric is far from the metal of the electrodes. There are two methods of building up a quenched gap, shown in Fig. 19 (*a*) and (*b*).



Internal Spark Gap. (*a*) *External Spark Gap.* (*b*)

FIG. 19.

In Fig. 19 (*a*), the gap consists of a number of metal discs with circular grooves cut in them. The discs are separated by washers of mica, about 0·2 mm. in thickness, which are inserted between the discs at the outer edges and extend a little way across the circular grooves. The sparking surface in the centre of the discs is thus completely shut off from the surrounding air.

In Fig. 19 (*b*), the mica washers are placed between the electrodes inside, and just extending into, the space between the circular grooves. The sparking takes place between the edges of the electrodes. This is the type used in the Service.

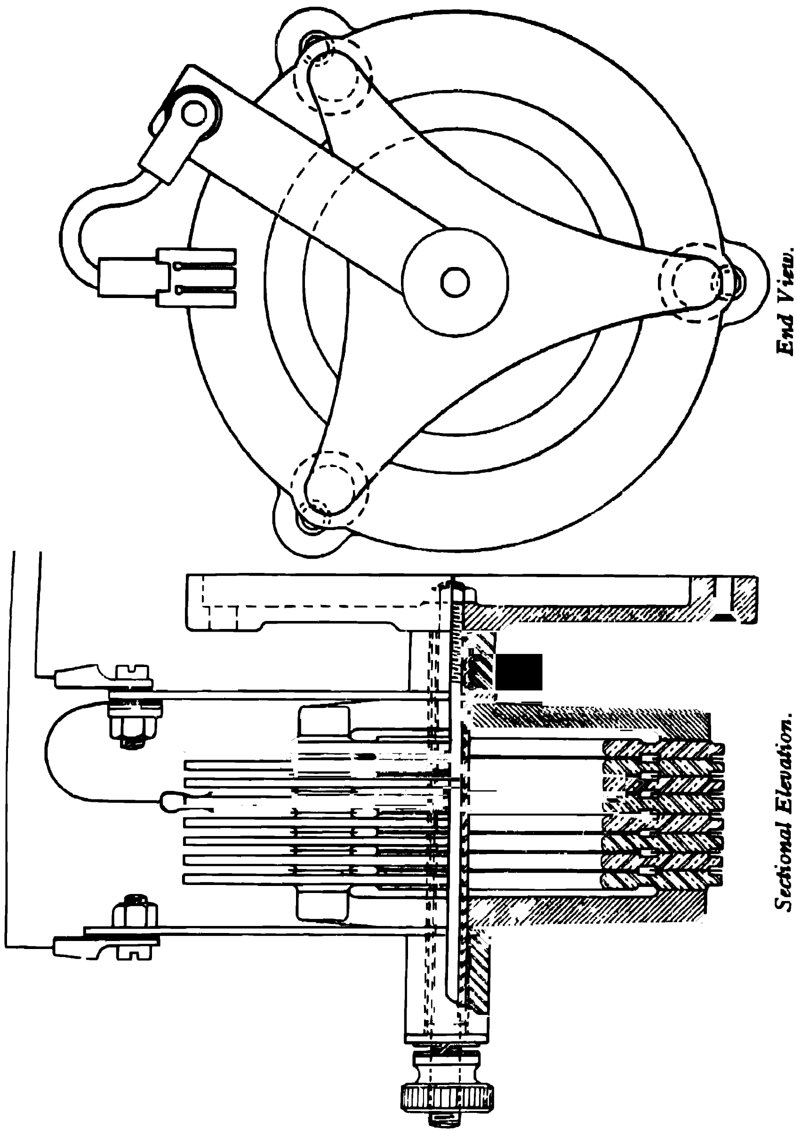
In both cases the spark is rapidly extinguished because of the adequate cooling properties of the gap, and because the mutual repulsion of the ions in the ionised air between the electrodes forces the spark to the outer edges of the sparking surface, where it becomes lengthened across the circular groove. In the " internal " gap it is probable that the removal of the oxygen in the gap by oxidation of the electrodes is an important factor in assisting its quenching properties.

Fig. 20 shows a complete quenched spark gap as used in the Service. The number of single gaps used in series can be varied by short-circuiting the others by a spring clip, which is attached to one terminal of the spark gap by a length of conducting wire and can be fastened to any one of the radiating fins. The number is varied according to the condenser break-down voltage, the voltage for each individual gap being about a thousand volts.

29. **The Charging Circuit Complete.**—We are now in a position to discuss broadly the general requirements of a Spark Transmitting installation (*see* Fig. 22).

Power Supply.—The alternating current will be supplied from either a motor alternator or a rotary converter.

Duplicate machines are generally supplied in H.M. ships, placed in separate watertight compartments, and fed through duplicate starters from either side of the ring main system.



Quenched Gap.

FIG. 20

Normally, the machine in use should be fed from the section of the ring main on the disengaged side in action.

With both motor alternators and rotary converters a **Starter** will be necessary, with a no-volt-release coil, and some arrangement to prevent an excess current being taken from the mains. This may take the form of a fuse in the lead from one main, or of an overload release coil on the starter.

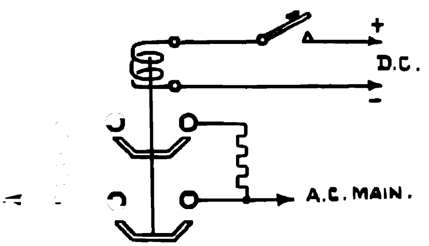
A Motor Field Regulator will be provided for varying the speed, and therefore the alternating requency. (As previously explained, the voltage of the alternating current supplied by a **rotary converter** will always remain about 65 per cent. of the supply voltage, whatever the speed.

An **A.C. Ammeter** and a **Frequency Meter** will also be provided. With a motor alternator, a **D.C. Ammeter** and an **A.C. Voltmeter** will be supplied ; also an **Alternator Field Regulator,** in order to vary the voltage of the alternating current.

From the slip rings of the machine in use come the A.C. mains, in series with which we shall have the Signalling Key, an Impedance Coil, a safety arrangement of some sort, and, finally, the primary of the step-up transformer.

30. **Signalling Key.**—If the alternating current is small and its voltage is low, signalling may be effected by making and breaking the alternating current circuit by means of a hand key.

If, however, the current is too great, or the voltage of a dangerously high value, a **Magnetic Key** will be used. This consists of a single-pole break between two contacts, which is short-circuited by a moving contact. The key is energised by direct current being supplied to the bobbin of a solenoid when a hand key is pressed, as illustrated in Fig. 21.



Magnetic Key.

FIG. 21.

When the low-tension current is too great to be interrupted conveniently, the magnetic key may be placed in the high-tension circuit, but special precautions must then be taken about its insulation.

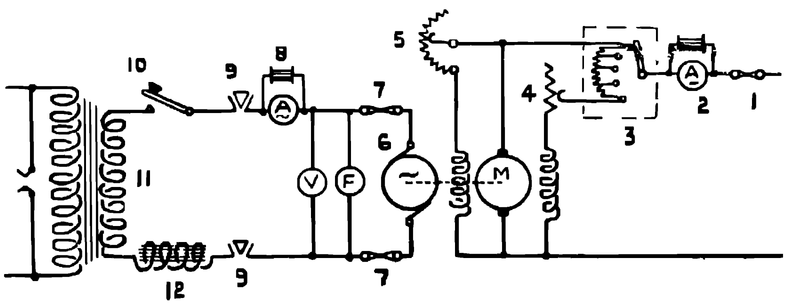
The drawback of this arrangement is that, even when the key is not pressed, the secondary terminals of the transformer are alive.

It is used in certain high-power shore stations.

31. **Safety Arrangements.**—It is necessary to prevent the operator from accidentally receiving a fatal shock.

For a shock to prove fatal, about one-eighth to one-sixteenth of an ampere of direct or low frequency alternating current will be necessary, one ampere being certainly enough to kill a man ; far heavier currents at high frequency, however, can be safely withstood.

The resistance of the human body varies in different individuals and with the degree of moisture of the skin. The greater part of the body resistance is in the skin and, normally, the resistance of a man from one hand to the other is about 20,000 ohms ; if the subject is in a perspiring state—a common occurrence in the tropics—the resistance will be reduced very much below this figure. It is also obvious that the resistance presented will be inversely proportional to the area of skin in



Typical Low-tension Circuit.

FIG. 22.

contact with the live terminals ; for example, a very much worse shock will be experienced if the live terminals are firmly grasped with the hands, or held in a pair of pliers, than if they are only touched with the finger tips.

Any voltage above 1,250 volts direct or low-frequency supply will probably prove fatal, and voltages of 500 and above are dangerous. It is sometimes customary to consider that sources of supply at pressures of the order of 220 volts can be touched with impunity ; this is a great mistake, for, with good contact and wet hands 220 volts can be—and has been—fatal. Moreover, should

the skin become broken by (say), scratching against 220 volt terminal contacts, the resistance becomes so low that a serious electrical burn may be produced, which might take several months to heal and even produce fatal shock effects.

In modern practice, it is customary to enclose all of the dangerous components within a cage, arranging that the cage cannot be opened without breaking the supply circuits at one point at least. It is better if both mains are broken, one by each of two cage doors. Thus, if the key is pressed when either cage door is open, nothing will happen.

32. **Low-tension Circuit.**—Fig. 22 represents a typical arrangement of a Low-tension Circuit.

1 is a fuse in the positive D.C. main.
2 is a D.C. Ammeter with its shunt.
3 is a starter for the motor of the Motor Alternator (6).
4 is the Motor Field Regulator.
5 is the Alternator Field Regulator.
6 is the Motor Alternator. (If a Rotary Converter is supplied in lieu, then the Alternator Field Regulator (5) and field magnet winding will be omitted.)
7/7 are A.C. fuses, one in each A.C. main.
8 is the A.C. Ammeter with its shunt.
V and F are, respectively, the A.C. Voltmeter and Frequency Meter.
9/9 are the breaks in the low-tension A C. circuit, completed when the cage doors are closed.
10 is the hand key for signalling. If the A.C. current is large, this is replaced by a magnetic key.
11 is the Transformer.
12 is the Impedance Coil.

If power is obtained from a Rotary Converter, then the A.C. Ammeter and Voltmeter may be omitted.

33. **High-tension Circuit.**—The high-tension circuit comprises the secondary of the transformer, the leads to the primary oscillating circuit, and the oscillating circuit itself, made up of primary condenser, primary inductance and spark gap.

In this circuit it is necessary to protect the transformer and condenser against excessive strains on their insulation, which may arise in several ways.

These risks, and the methods of guarding against them, are as follows :—

34. **Safety Gaps.**—The transformer is protected against any excessive rise in its own voltage by having safety gaps fitted across its terminals.

The gap is so arranged that if the terminal voltage of the secondary rises above its normal working limit, then an A.C. metallic arc will form across the gap.

Typical safety gaps are shown in Fig. 23.

Fig. 23 (*a*) consists of two bent pieces of copper wire. Fig. 23 (*b*) consists of two brass discs, mounted eccentrically so that the gap length can be adjusted by rotating them.

(*a*) (*b*)

Safety Gaps.

FIG. 23.

The arc will rise, on account of its own heat, sliding along the wires. The higher it rises the longer becomes the gap which it must bridge, so that it automatically breaks down.

The gap must be kept set at the exact distance laid down in the Handbook for the set in question.

In the same way, the transmitting condenser is fitted with safety gaps to prevent it being punctured by any abnormal rise of voltage.

It is commonly the practice, where several sections of a condenser are employed in series, to earth the centre point of the central sections.

This equalises the capacity to earth of each element of the condenser.

35. **Back Oscillations.**—When the condenser discharges across the spark gap it is quite possible that some of the high-frequency oscillatory current, instead of flowing across the spark gap, will try to flow back through the secondary of the transformer.

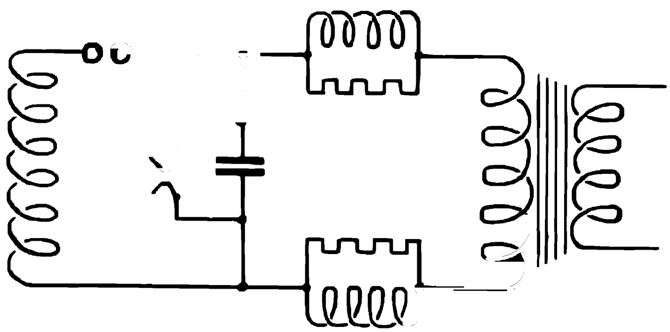
This is a form of trouble known as " Back Oscillations."

If this high-frequency oscillation is applied directly to the secondary turns of the transformer, it will set up a big inductive voltage to earth across the end turns, and the insulation of the transformer might easily be broken down.

To prevent this, coils of wire, termed " Protecting Coils," of about 200 to 300 mics inductance, are sometimes placed in series with each main (Fig. 24).

The back E.M.F. of these coils to the **low-frequency charging current** is very small indeed, nor does the addition of a few mics disturb the resonance of the charging circuit.

On the other hand, when the **high-frequency discharge** tries to send some of its current back along the high-tension mains, the reactance of the protecting coils becomes so large that they practically form insulators to high-frequency currents. (The insulation of these coils may be punctured, but they can easily be re-wound on board.)



Protecting Coils and Resistances.

FIG. 24.

Now, protecting coils alone, under certain circumstances, have proved worse than useless.

Being mounted on a baseboard, the two coils have a certain small capacity effect between them ; also, each coil has a certain self-capacity of its own.

This capacity, combined with the inductance of the two coils, forms an oscillatory circuit which may happen to have the same LC value, and therefore the same natural frequency as that of the primary oscillator.

Should this happen, then resonant currents will be set up in the protecting coils, and high-frequency currents will be impressed on the end turns of the transformer windings, with the result that those turns will have their insulation punctured.

To avoid this, the protecting coils are shunted with non-inductive resistances in the shape of **Carbon Rods**, upon which the energy of the high-frequency currents is expended.

The high-frequency current sets up a big P.D. across the inductance, which consequently tries to force a big current through the resistance, and the back oscillation is damped out.

Fig. 25 represents the high tension circuit of a modern spark attachment. The oscillatory circuit is connected to the secondaries of the transformer through a resistance " R " ; connected across the output of the transformer are two condensers " C." The " protecting condensers " C serve to by-pass any R/F oscillations which may be fed backwards, thereby preventing the building up of high R/F voltages across the secondary of the transformer. The resistance R is of the order

of 600 ohms and is there in order to prevent the two protecting condensers from acting as integral components of the main oscillatory circuit ; the ratio of R to the equivalent reactance of condensers C, determines the extent of their influence on the frequency.

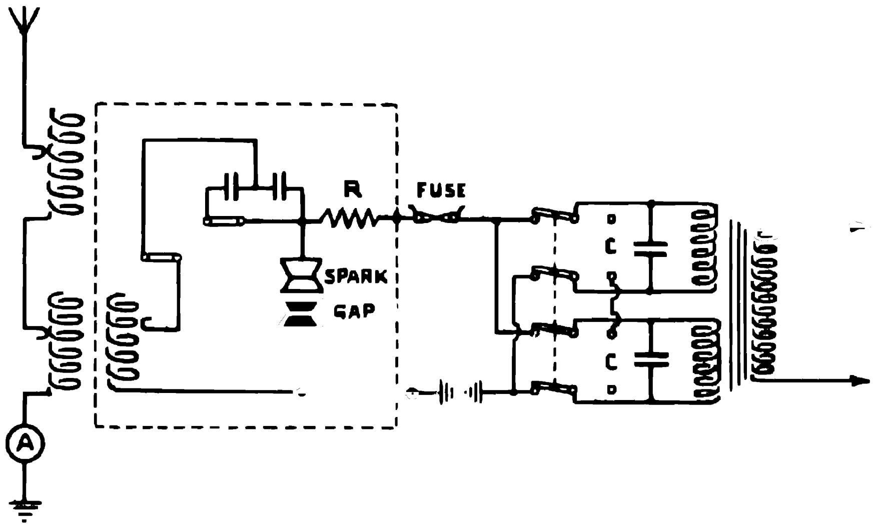


FIG. 25.

36. **The Design of the Oscillatory Circuit.**—The oscillatory circuit must be arranged so as to provide a certain range of LC values, according to the purpose for which the set is designed.

For example, a set might be required to give a range of all LC values between 50 and 1,200, which would give a range of frequencies between 676 kc/s. and 140 kc/s.

This might be arranged by having a condenser of 50 jars and an inductance which could be varied between 1 mic and 24 mics, but this would be bad electrical practice for the following reason :—

The energy stored in a condenser of C farads when charged up to a discharging voltage of V volts is $\frac{1}{2}CV^2$ joules.

From this it follows that, the larger the condenser, the smaller the voltage required to store a given amount of energy. (*See* para. 22.)

A high condenser voltage would mean a long spark gap, which would have a heavy damping effect. Also, the higher the voltage, the greater do insulation difficulties become.

Consequently we want to use the largest possible condenser and the smallest possible inductance.

A much better arrangement would be to provide a condenser composed of two elements of 100 jars each, which could be joined in series, giving a capacity of 50 jars, or in parallel, giving 200 jars. This arrangement is generally used in practice.

A primary inductance which could be varied between one and six mics would give a range of LC values from 50 to 300 in the series position, and from 200 to 1,200 in the parallel position.

37. **The Condenser.**—The condenser must be made with adequate dielectric strength to stand the greatest sparking voltage that is likely to be used. For instance, if the condenser referred to above is required to stand an eight-millimetre spark, and if one sheet of the dielectric used in it

will only stand two millimetres safely, it would be necessary to make up the two elements of 50 jars each by joining four sections of 200 jars each permanently in series. It would therefore be composed as follows :—

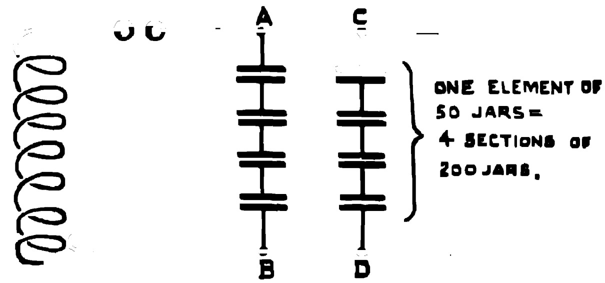


FIG. 26.

To join the two elements in series, connect terminals B and C together.

To join the elements in parallel, connect A to C and B to D.

The dielectrics generally used in the Service for spark transmitting condensers are ebonite, glass or mica.

The elements and sections composing the condenser are contained in an iron tank, which is kept brimful of oil, to prevent brushing over the plate edges, and to keep the condenser cool.

However efficient a condenser is made, certain losses in it are unavoidable. These losses were referred to in Vol. I under the general heading of " Hysteresis Losses."

When a condenser is being charged up and discharged, it gradually gets hotter and hotter as a result of these losses.

Arrangements have to be made for this heat to be radiated away, and also for the oil to expand without forcing its way under the edge of the lid.

38. **The Primary Inductance.**—This is a coil of copper tubing or flat copper strip, of large surface area, in order to obtain low resistance to high-frequency currents with a given required inductance and mechanical rigidity. It must have a surface of adequate area for the maximum oscillatory currents it will be required to carry.

The turns must be spaced sufficiently far apart to prevent sparking over between adjacent turns.

An adjustable connection, of very low resistance, must be provided so that the inductance in circuit can be varied gradually from the minimum to the maximum value, in order to give any required LC value between the limits for which the circuit is designed.

39. **The Aerial Circuit.**—This circuit, illustrated in Fig. 27, comprises the following :—

Aerial, Feeders and Deck Insulator, which are fully dealt with in Section " R."

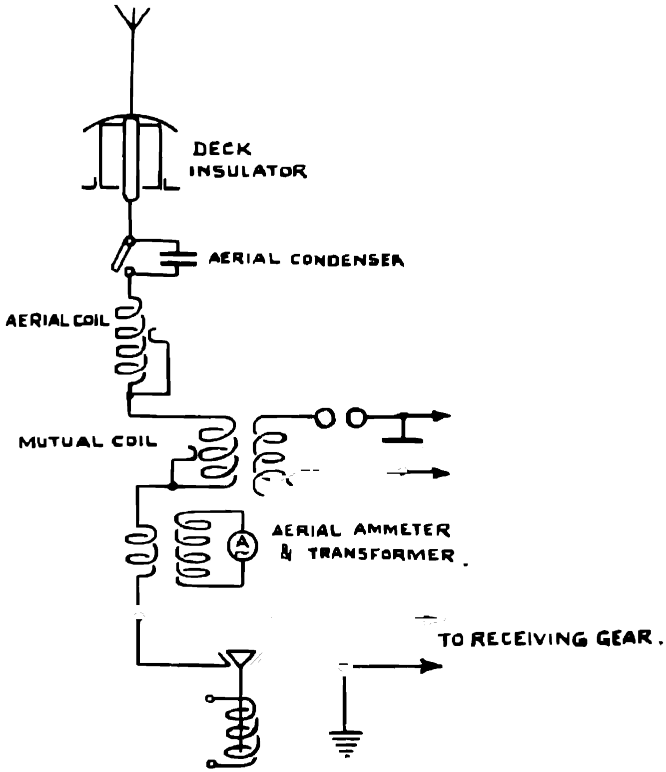
(*a*) **The Aerial Coil.**—The aerial coil is provided in order to increase the LC value of the aerial circuit when transmitting waves longer than the fundamental wave of the aerial.

In spark transmitting circuits it is made of stout copper wire, the turns of which are wound on a former of insulating material, and spaced sufficiently wide apart to prevent sparking over between adjacent turns.

An important point is whether the idle turns of the aerial coil, *i.e.*, those not required for the particular frequency in use, shall be short-circuited, as in Fig. 28 (*a*), or left on open circuit, as in Fig. 28 (*b*). This question has already been discussed in Vol. I.

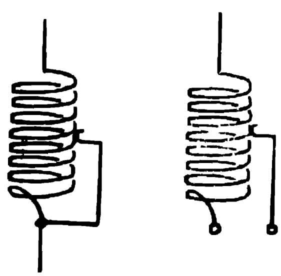
The general practice is to short-circuit the idle turns of coils used in spark transmitting sets.

When making a tuning connection on the aerial coil, it is important to remember that all the current of a received signal has to pass through it, and therefore a thoroughly clean and tight connection should be made.



Typical Aerial Circuit.

FIG. 27.



(*a*) (*b*)

FIG. 28.

(*b*) **The Mutual Coil.**—The mutual coil is provided in order to transfer into the aerial circuit the high-frequency oscillations generated in the primary.

Its distance from the primary is made adjustable in order to allow the coupling to be varied. In some circuits its inductance is also made adjustable.

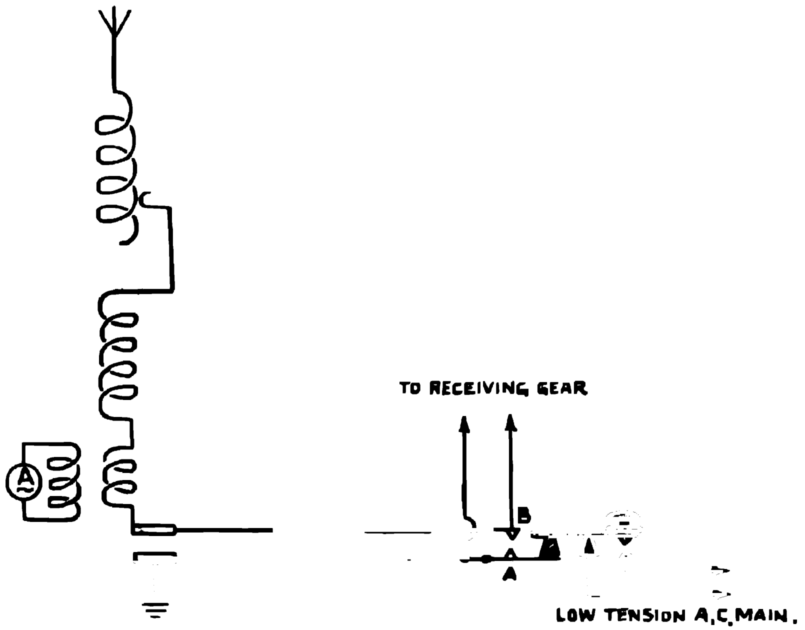
The loosest possible coupling consistent with giving readable signals to the receiving station should always be used.

It is better to use a long spark and a loose coupling than a short spark and a tight coupling.

(*c*) **The Aerial Condenser.**—For transmitting waves shorter than the fundamental wave of the aerial, a **series condenser** is generally used, being short-circuited with a link when not required.

40. (*d*) **The Aerial Ammeter.**—In order to enable an operator to tell that his circuits are correct, a hot-wire ammeter, termed the "Aerial Ammeter," is provided.

If joined directly in series with the aerial, it would have to be unduly large to carry the full aerial current in most aerial circuits. It is therefore usually joined across the secondary of a toroidal transformer.



"*Listening Through*" *Device*

FIG. 29.

41. (*e*) **Send-Receive Contact.**—This consists of a make-and-break in the earth lead, across which the receiving gear is joined.

When the transmitting key is pressed this break is short-circuited, thus short-circuiting the receiving gear, and connecting the mutual coil to earth for transmitting purposes.

When the transmitting key is released the break is opened, thus putting the receiving circuit in series with the aerial.

This allows for what is termed "listening through," *i.e.*, listening between the Morse signals of one's own message to see whether anyone else is transmitting at the same time.

The requirements of the break are :—

(a) It must make before, and break after, the low-tension charging circuit, to obviate the risk of sparking into the receiving gear.

(b) It must either be very close to the earth connection or be joined to it by a non-inductive lead. If there were a long inductive lead between it and earth the voltage across it would be considerable, and the break would have to be a long one to prevent sparking taking place across it as it was made and broken.

A suitable arrangement is illustrated in Fig. 29. Here were have a hand key, fitted with two " back contacts," A and B. Contact A is carried on a flexible springy piece of copper.

A and B are connected to aerial and earth by a non-inductive lead of concentric cable, and across them is joined the receiving circuit. When the key is at rest, an ebonite thimble on the toe of the key is bearing on the end of the copper strip that carries contact A. The aerial is then connected to the receiving gear.

When the key is pressed, however, its first motion allows A and B to make contact, and thus to complete the aerial circuit to earth, and to short-circuit the receiving gear before the "heel" contacts make.

Its further motion closes its " heel " contacts, and completes the low-tension A.C. circuit.

EXAMINATION QUESTIONS ON SPARK TRANSMITTERS.

1. Sketch the circuit of a simple spark transmitter. Explain the action of the protecting coils, the carbon rod resistances, and the safety gaps.

   (Qualifying for P.M.G. II Cert. H.M. Signal School, 1934.)

2. Describe and explain the action of a synchronous spark system of wireless telegraph transmission.

   (I.E.E., Oct., 1927.)

3. Draw a circuit diagram of a rotary spark transmitter of about 1½ kW. input, assuming that a D.C. supply is available. Indicate the usual switches, fuses, and transmitting key.

   (C. & G., I., 1928.)

4. Describe, with sketches, three types of spark gap used on spark transmitters and state the advantages and disadvantages of each type.

   (C. & G., I., 1930.)

5. Describe, with a diagram, the construction and working of a ½ kW. spark transmitter suitable for marine use. What are the advantages of spark sets over I.C.W. sets for marine emergency purposes?

   (C. & G. Preliminary, 1934.)

THERMIONIC VALVES.

1. **Historical Introduction. The Fleming Valve.**—In the eighties of last century, Thomas Edison, of the United States of America, discovered that an electric current could be made to pass across the empty space between the hot filament of an electric lamp and another metallic conductor, or collector electrode, contained in the same evacuated glass bulb. He made the far reaching observation that the **current travelled in one direction only.** No explanation of the phenomenon was at first available, and it was not until after 1899 that J. J. Thompson's discovery of the "electron" gave a clue to the nature of the "Edison effect." It had been clear to J. A. Fleming that the current was due to something travelling from the hot wire to the collecting plate, and after Thompson's discovery the suggestion became irresistible that the hot filament was actually emitting unit particles of negative electricity, now called **electrons.**

In 1904, J. A. Fleming had the vision to realise that the phenomenon of "thermionic emission" had a commercial application for the detection of wireless waves; he saw the potentialities of an electric "valve," a device passing current in one direction and prohibiting it in the opposite direction. The early wireless experimenters sought diligently for something which would react to the minute alternating potentials produced in a receiver of wireless waves. The frequencies in use were far beyond the powers of any mechanically operated indicators; the diaphragm of a telephone receiver, for example, can vibrate at audible frequencies from a few hundreds to a few thousands of cycles per second, but is unaffected by the frequencies of the order of hundreds of thousands of cycles per second which characterise radio waves. Fleming's valve, by passing the current in one direction only, gave a *D.C. output from a radio frequency A.C input,* thus constituting an "indicator" of wireless waves. The D.C. current could be made to affect the diaphragm of a telephone receiver, an audible note being produced if arrangements were made to interrupt the radio waves at an audible frequency.

Fleming's two-electrode valve, operating in this manner, became a "detector" of E.M. waves, the process of **detection** being dependent upon the "rectification" of the R/F alternating currents set up by the incoming signal. In its original form, the Fleming valve had a cylindrical electrode made of aluminium, and a hairpin carbon filament which emitted the electrons. Subsequent developments have added emphasis to the importance of Fleming's application of the Edison effect. It is not an over-statement to say that the rate of evolution of the valve has been the factor controlling the advances in wireless technique.

The emitter of electrons in a valve is called the "cathode"; the collecting electrode is usually called the "anode," but is sometimes called the "plate." A valve with these two electrodes, **cathode** and **anode,** is called a **diode.**

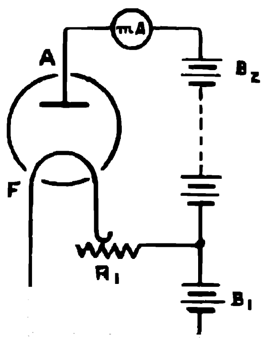
2. **Multi-Electrode Valves.**—In 1907 Dr. Lee de Forest took out a patent in America for a valve like Fleming's, but with a third electrode, consisting of a "grid" of wires, interposed between the cathode and the anode. He found that the potential of this **grid** had a remarkable control over the anode current, a factor which enormously increased the potentialities of the valve. These will be described in later paragraphs and sections of this work, but it may be said here that the three-electrode valve, or **triode** as it is generally called, can be used as a detector of oscillations, as an amplifier, and as a generator of oscillations.

In recent years valves have been developed with still more electrodes. The additional electrodes give rise to various properties which are applied for particular purposes. The words "diode, triode tetrode, pentode, hexode, heptode, octode," etc., are used to indicate two-electrode, three-electrode, four-electrode, etc., valves respectively. All of these valves, however, are to be regarded as a direct result of de Forest's discovery that the anode current may be controlled by the potential of an interposed grid.

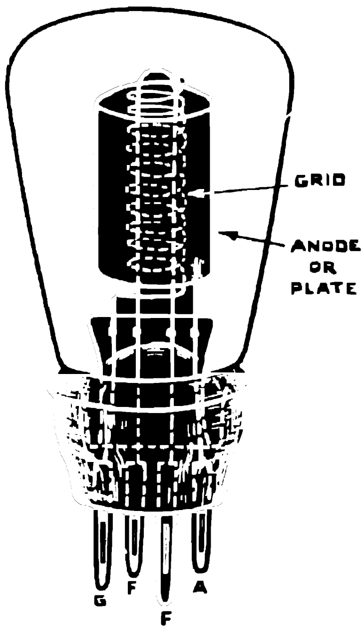
3. **Construction of Thermionic Valves.**—A valve consists, externally, of an envelope or bulb of glass or other suitable material; the electrodes are placed within the bulb and the whole is then usually highly evacuated. The materials of which the electrodes are made are deter-

mined by the functions they perform, and the use to which the valve is put. For example, valves now used for transmitting purposes are larger and are subjected to much higher voltages than those used for receiving purposes, detectors and amplifiers, etc. ; the essential features remain the same.

4. **Thermionic Emission. The Diode Circuit. The H.T. Battery.**—Fig. 1 (*a*) represents a diode, and it is shown here attached to two batteries, one of which provides the necessary current to heat the filament, and the other is joined between anode and cathode for a reason to be explained later.



(a)



(b)

FIG. 1.

It is worth while to reflect briefly upon the mechanism of the process known as "thermionic emission," *i.e.*, the ejection of electrons from a hot body.

In the explanation of the electron theory in Vol. I, it is stated that the mobile electrons present in a substance which is an electrical conductor, are continually moving about from one atom to another and that, if the substance is heated, the velocity of the electrons is increased. If the temperature is raised sufficiently, it is believed that, in certain substances, the electrons are agitated to such an extent that some of them leave the surface of the material. This *evaporation* of the electrons from the surface of a body is very similar to the evaporation of molecules of liquid at a surface exposed to air, to the turning of water into water vapour, etc. ; similar quantitative laws have been found to apply. The phenomenon is known by the term **thermionic emission.**

If there is no external field acting on the escaped electrons, they return to the parent substance, which their absence has left positively charged. Thus, any conductor when heated above a certain temperature is surrounded by a cloud of electrons (the "**space charge**") which are continually being shot out of it, and attracted back into it. This emitting property of a hot metal is the basis of the thermionic current in the valve. The hot metal is the filament or cathode, and is made of a material which, when raised to a suitable temperature less than its melting point, gives adequate electron emission for the purpose to which the valve is to be put.

The method used to heat the filament is to pass a current through it. In valve theory we shall be much concerned with batteries connected between the various electrodes, and it is important here to grasp clearly the idea that the battery which is connected across the ends of the filament is there simply because it is the most convenient way of raising the temperature ; the heating current applied to the filament is entirely distinct from the emission current, or the currents that flow from one electrode to another.

A low voltage battery of 2, 4 or 6 volts is used for heating the filament of a small directly heated receiving valve, battery B_1 of Fig. 1 (*a*). In certain cases an adjustable resistance R_1 is connected in series with the filament to regulate the heating current flowing. Sometimes filaments may be heated by alternating current, or the electron emitting surface may be a cathode which is indirectly heated.

If another electrode, A of Fig. 1 (*a*) is placed close to the filament and raised to a positive potential with respect to the filament, by means of battery B_2, some of the emitted electrons will experience an accelerating force and travel towards the anode.

If no battery B_2 were used, there would still be a small anode current due to the "Edison effect." The electrons have a certain velocity on breaking through the surface tension of the filament, and it is possible for an anode to collect those which are driven on to it by their velocity, no difference in potential being necessary. The effect is a very small one, and will not be further considered.

In general, the anode is made considerably positive to the filament. The anode voltage of a *receiving valve* varies from 50 to 500 volts, and is generally provided by joining a high tension battery between the outside terminal of the anode and one of the terminals of the filament, or the cathode in the case of indirectly heated valves. In valve work, the conventional datum of reference to which potentials are referred is the L.T. negative terminal, or the cathode in the case of indirectly heated valves. The anode voltage of *transmitting valves* may be from a few hundred to ten or twenty thousand volts, and may be provided from high tension D.C. machines, or from alternators and rectifiers.

When the negative electrons arrive at the anode, they pass along the conducting wires from the battery and back to the filament. There is thus an electrical current round the circuit comprising valve, battery and leads. Attention may be directed, again, to the difference between the direction of the actual flow of electrons around the circuit and the flow of electrical current as conventionally assumed. The electrons travel through the valve from negative to positive, whereas the flow of an electrical current has always been assumed to be from the positive terminal of a source of current to the negative one. In ordinary electrical parlance the anode current is spoken of as "flowing from the positive terminal of the battery, through the valve, and back to the battery at the negative terminal." In the case of receiving valves the anode current is generally of the order of a few milliamperes ; in a large transmitting valve it may be many amperes.

Fig. 1 (*b*) explains the construction and disposition of the electrodes in the case of a triode valve.

It is proposed to treat, in slightly greater detail, the construction of the various elements composing the valve.

5. **Construction of the Cathode.**—Various materials are used in the construction of cathodes, both of the directly and indirectly heated types. Experiments show that different materials emit different quantities of electrons at the same temperature, and that at ordinary temperatures there is no measurable emission from the surface of any material. Thermionic emission is a surface effect, and it depends on the physical nature of the surface as well as on the temperature.

There are three principal classes of electron emitting materials now used in valve construction :—

(*a*) Pure tungsten wire, operating at temperatures between 2,400° and 2,500° absolute.
(*b*) Thoriated tungsten wire, operating at temperatures between 1,800° and 1,900° absolute.
(*c*) Certain "oxides," operating at temperatures between 1,100° and 1,300° absolute.

The absolute scale of temperature, which is that generally adopted for filament temperatures, has its zero at − 273° C. The degree intervals are the same on both scales, so that a reading on the absolute scale is obtained by adding 273 to the corresponding reading on the Centigrade scale.

6. **Tungsten Filaments.**—Tungsten is a metallic element, having a melting point at 3,400° C. This is a value higher than that found for any other of the common metals, and since the emission of light increases with temperature, it is obvious that the metal which can be produced in a suitable form and can also stand the highest temperature, is the most suitable one to use as a lamp filament. It is precisely this same fundamental property which accounts for the use of tungsten both as an electron emitter, and as a source of light in ordinary lamps.

In order that the filament of a lamp should have a long life, it was found to be essential to exhaust the bulb to the highest possible vacuum. The same condition was soon found to be desirable for preserving good and uniform conditions of electron emission. Thus, in two important features, the valve industry profited by the experience of the lamp industry, and as similar methods of

assembly, using identical machinery, could be employed to a very large extent for both, the two industries have been closely associated for research and production purposes. In recent years, however, it has been found possible to operate tungsten filaments at still higher temperatures, and so produce more light, by filling the bulbs with an inert gas, such as argon, after all traces of oxygen and water vapour have been removed. The majority of thermionic valves, on the other hand, still require the highest possible vacuum for efficiency, so that there is now not so much in common in the modern products of the two industries.

Tungsten filaments have been used in a wide range of diameters from about 0·05 mm. in receiving valves, up to 2 or 3 mm. in the largest transmitting valves ; the corresponding range of heating currents is from about half an ampere to some hundreds of amperes.

7. **Thoriated Tungsten Filaments.**—Thoriated tungsten wire is made by adding up to about 1 per cent. of thorium oxide (or thoria) to the tungsten wire material during manufacture. In the early days of electric lamps, it was found that pure tungsten filaments were liable to fracture suddenly, generally on switching the current off, and this was traced to a change in the crystalline nature of the material. It was found that the addition of a trace of thoria prevented this defect to a considerable extent, and it was during observations on valves having filaments constructed with thoriated lamp wire, that the enhanced electron emitting properties of this material were accidentally discovered. These filaments give a copious emission of electrons at a temperature some 600° C. lower than that normally used with bright tungsten filaments. Less heating current is required, and the filament glows less brightly.

According to modern theory, based on extensive researches, the thorium oxide is decomposed in the filament, and the pure metal thorium forms a layer on the surface of the tungsten, and is the factor mainly responsible for the enhanced emission of electrons. The tungsten wire merely serves as a mechanical support and conducting base for the coating of the filament.

The layer of thorium is not extremely robust and is easily removed by excessive temperature above the normal operating temperature of 1,900° absolute ; the coating of thorium evaporates, and, before long, the emission characteristic of tungsten is all that is available. If, however, the filament is run for several minutes at about 2,200° absolute, a new coating of thorium appears, and, on reverting to the normal operating current, the original emission will be obtained.

If traces of gas remain in the valve, the filament becomes subject to a bombardment by the positive ions formed by ionisation of the gas ; this also wears away the surface of the filament. The ionic bombardment of the filament is greatly intensified by high anode voltages and, for this reason, the thoriated filament is not suitable for use in high powered transmitting valves. Thoriated filaments are seldom subjected to more than about 150 anode volts.

Normal values of filament current range from 0·1 to 0·3 amps.

8. **Oxide Coated Filaments.**—The oxides of barium, strontium and calcium, are copious emitters of electrons at moderate red heat, a fact which was known before the Fleming valve was produced. For many years it had been known that a plentiful thermionic emission could be obtained at temperatures of about 1,100° absolute from a wire coated with a " dope " of any of those oxides. The first valves to be made with these filaments possessed very long lives and were very economical in battery power. Almost all of the ordinary receiving valves are now of this type.

It is now generally believed that the oxides are decomposed by the heating and electrical treatment to which they are subjected, and that the actual emission of electrons takes place from the pure metal,—*e.g.*, barium—set free by those changes ; the action appears to be similar to that of thorium in the thoriated tungsten filament.

Normal heating currents range from 0·075 amperes upwards. The application of too great a heating current involves the volatilisation of the active metal, and even the oxide coating itself If the cathode is not too extensively damaged in this manner, it can be reactivated by slightly overheating for a short period, as in the case of thoriated tungsten wire. There is also a similar limitation to the maximum anode voltage.

Since coated filaments are not heated so strongly as the other two types, they do not expand so much, and consequently it is possible to bring the grid wires of a triode much closer to them, and so to increase the control exercised by that electrode.

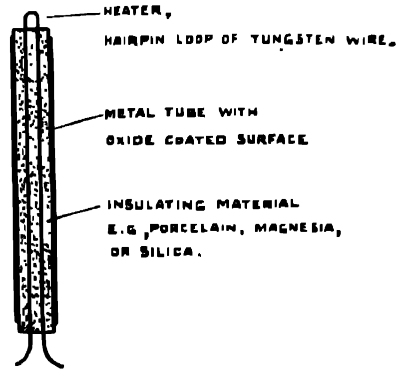
The core material most widely used for coated cathodes is nickel, though tungsten, molybdenum, and alloys of these and other metals are also employed. The question of cost is the prime consideration, and is only second in importance to the technical requirement that the metal core must retain the coating with tenacity.

The oxide coating may be applied in various ways. In one method, the active material is caused to settle on the filament wire after the latter has been assembled in the valve. A barium compound—barium azide—is incorporated in a "dope" painted on the inside of the anode. In the manufacturing process the anode is heated, the dope volatilises, and some of it settles on the filament. The latter is specially prepared and is heated slightly by a suitable current to enable the most effective adhesion of barium compound to take place.

9. **Indirectly Heated Cathodes.**—These were introduced to enable alternating current to be used with receiving valves, a feature of the modern "all-mains" receiver.

If an ordinary filament is supplied with alternating current at commercial frequencies, the fluctuations of voltage along its length produce corresponding fluctuations of the valve characteristics. The detector valve is particularly susceptible to this trouble, producing a hum associated with the frequency of the power supply. This is, of course, amplified by the following stages of the receiver.

The indirectly heated cathode overcomes the trouble. The filament itself is not the emitter, but provides the heat necessary to produce emission from an emitting surface which surrounds it. The heater is a stout hairpin tungsten filament, with a standardised rating of 4 volts, 1 ampere. It is coated with an insulating material such as aluminium oxide, or silica, and is inserted in a close fitting nickel cylinder—Fig. 2—on the outer surface of which the oxide coating has been sprayed. The coated nickel cylinder is the cathode proper, and there is, of course, an extra lead from this electrode which is connected to an additional pin in the base of the valve. The heater takes about half-a-minute to warm up the cathode to the emitting temperature, a point which has to be remembered when switching on receivers using these valves. On account of the high wattage of the heater, it is obvious that valves with indirectly heatded cathodes cannot be used in large numbers on batteries.



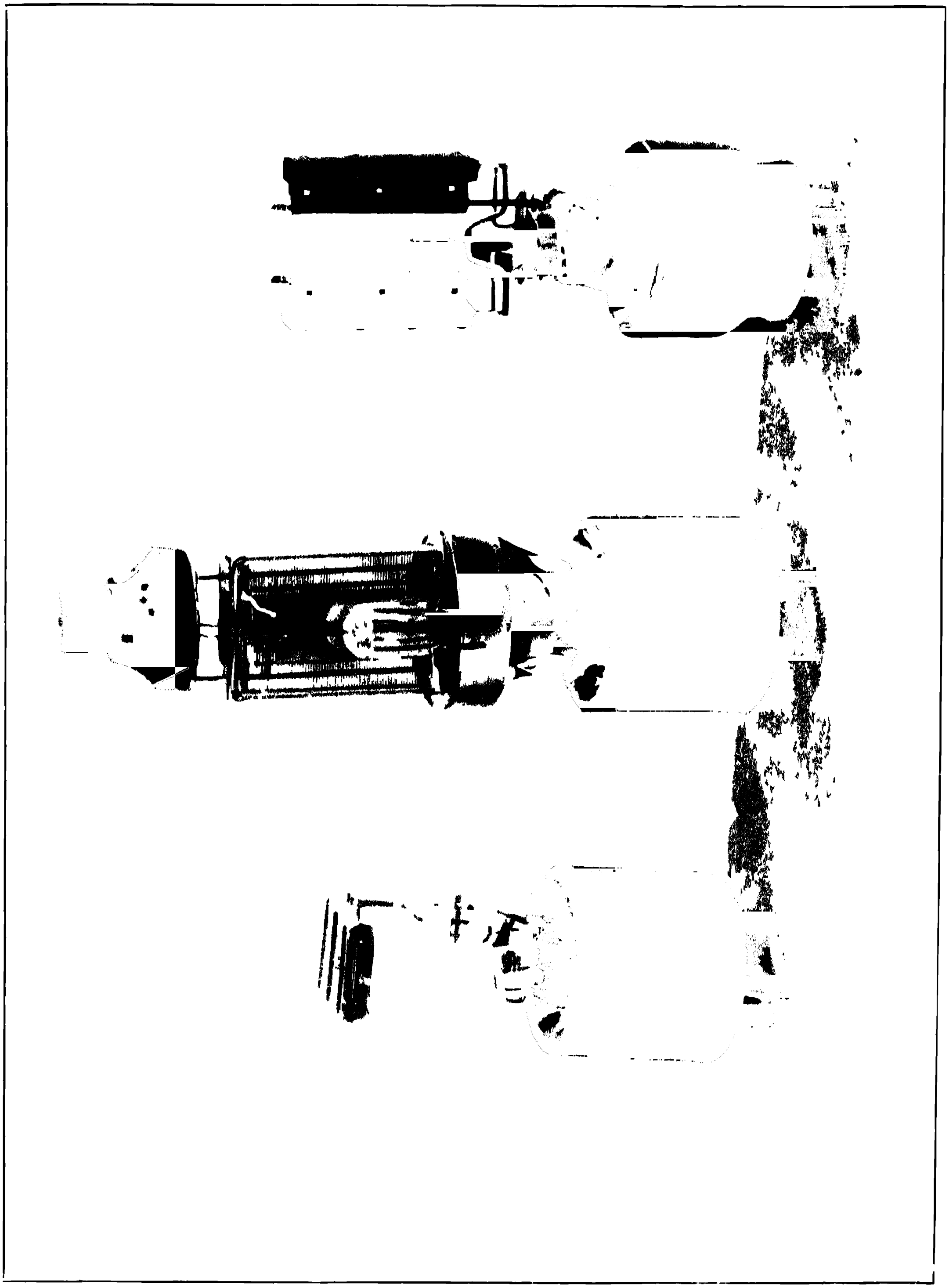
INDIRECTLY HEATED CATHODE SYSTEM.

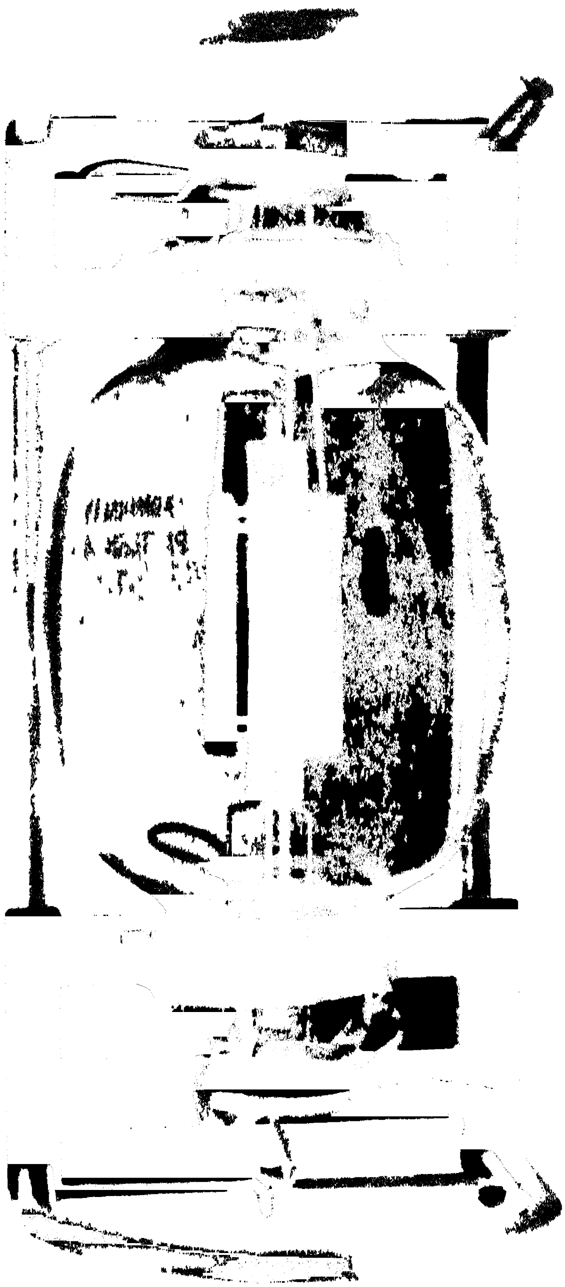
FIG. 2.

In counting the electrodes in a valve, the heater and cathode count as "one"; a valve with anode, grid and indirectly heated cathode is a "triode" and not a "tetrode."

10. **Materials for Grids and Anodes.**—These electrodes are made of various metals, nickel and molybdenum being the ones most widely used. In the case of anodes, it is essential to choose a metal which has a high enough melting point to withstand the heat generated, both in the process of manufacture and in its use, and the one which can be relied upon not to exude *occluded* gases which would spoil the vacuum of the valve. Nickel and molybdenum satisfy these requirements fairly well, but tantalum, zirconium, copper, and other metals, are also used in particular cases where their respective peculiar properties are of advantage.

More recently, carbon anodes have found an application in transmitting valves, when it is desired to increase the anode rating of the valve without providing cooling arrangements. There is the advantage that secondary emission (paragraph 24) is reduced, but against this must be set





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SECTION C B

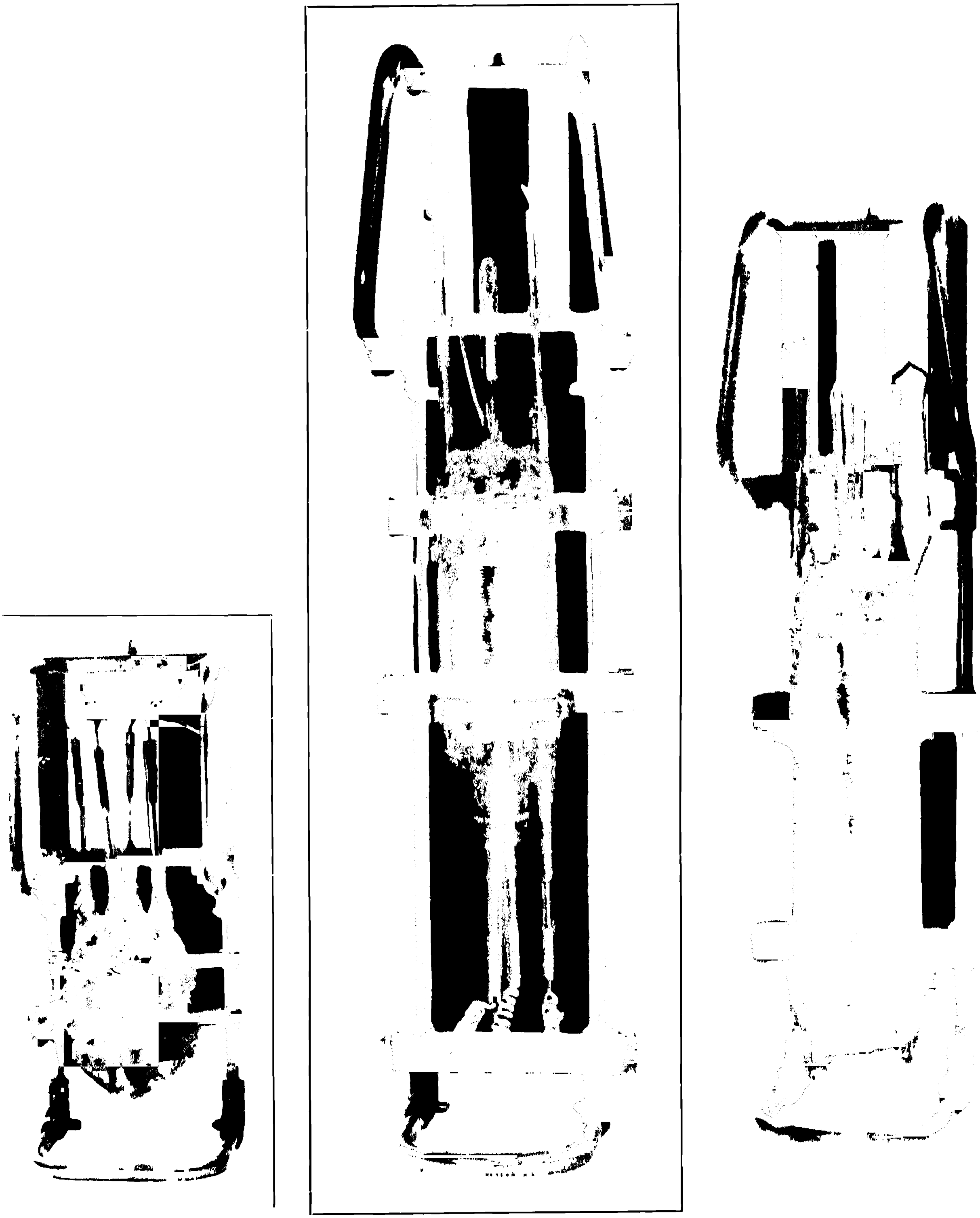
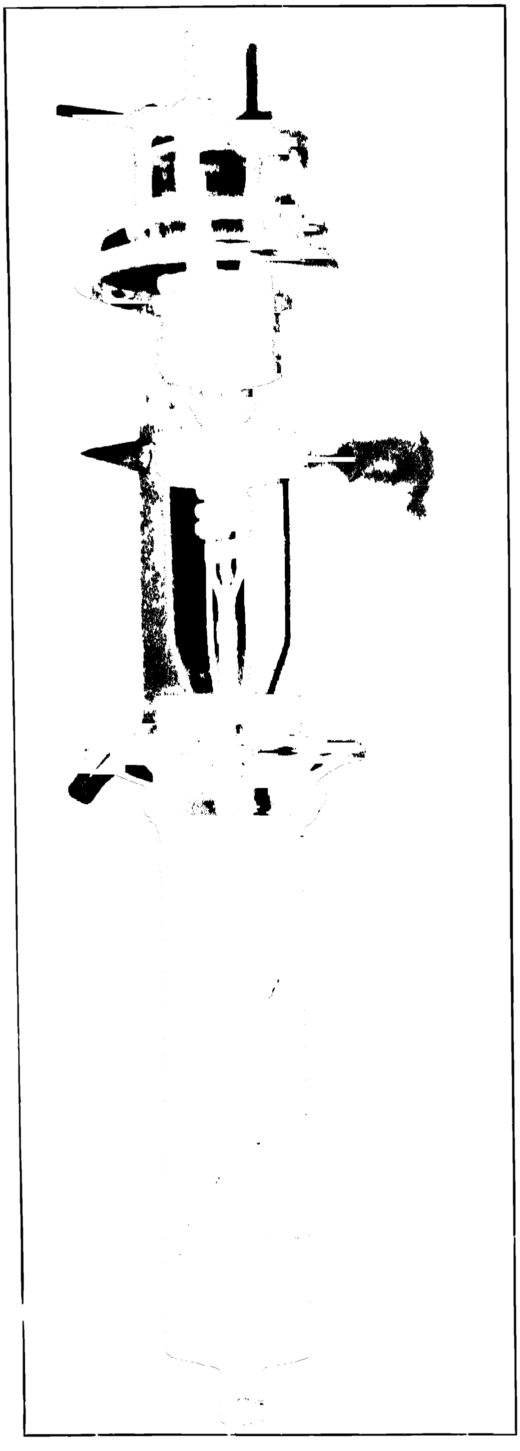
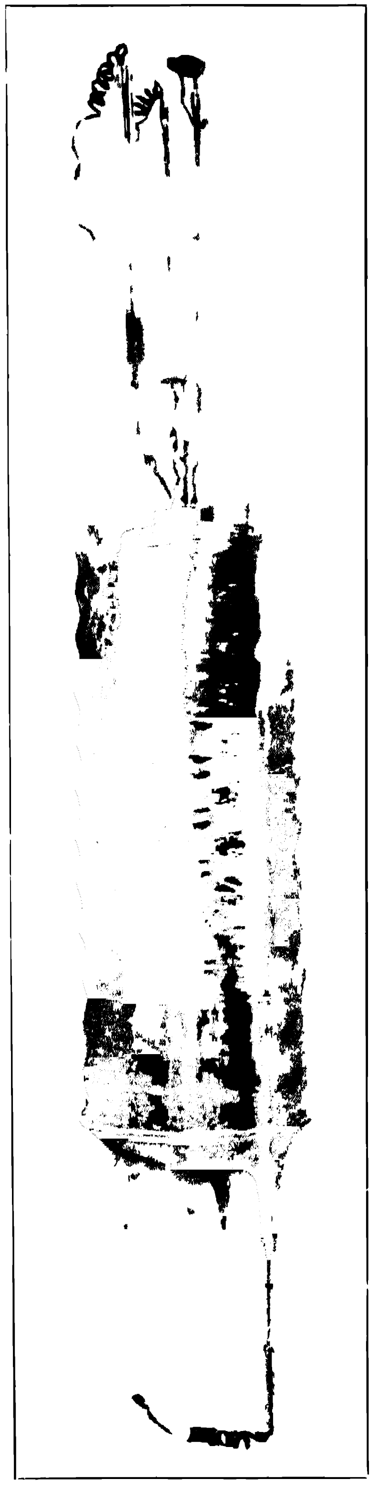


FIG. 5

SECTION "B"





the technical drawback that carbon anode valves are much more difficult to " outgas " than valves with metal anodes ; the carbon anodes occlude vast quantities of gas and make the evacuation a long process.

The high melting point of molybdenum makes it expensive to produce sheets of the metal ; it is therefore the practice to weave cylinders of molybdenum strip, forming anodes of basket-like appearance. The strip is about 1·5 mm. wide and is produced by rolling thin wire, a comparatively cheap operation.

During the process of manufacture, every attempt is made to eliminate the gases occluded in the electrodes. These are heated as much as possible after assembly, and the evolved gases pumped away. When the vacuum falls, a valve is said to become " soft " ; this may sometimes occur when a valve is in use, and may be due to the heating of the anode to a temperature higher than that employed during manufacture, resulting in the evolution of further small quantities of occluded gas.

Fig. 3 illustrates some of the various forms of electrode construction used in small receiving valves ; it shows the horizontal and vertical method of electrode assembly, the right-hand valve being a double diode.

11. **Anode Rating of Valves.**—When electrons are driven across the space from the filament to the anode, the electronic bombardment produces heat in exactly the same manner as that produced when bullets hit a target. If 1 ampere is flowing, and the P.D. is 1,000 volts, then the rate at which energy is converted into heat in the anode is 1 ampere $\times$ 1,000 volts = 1 kW. Valves have to be designed with anodes capable of dealing with the heat generated in this manner. In small receiving valves a few milliamperes flow under a P.D. of 100 volts or so, and the anodes are only required to deal with a fraction of a watt or perhaps a few watts. But in large transmitting valves, the power which is wasted as heat may be many kilowatts, and it is necessary that the shape, size and nature of the anode should enable this heat to be dissipated safely without injury to any parts of the valve.

If a valve is overrun, there is the risk of actually melting the anode. This sometimes happens in small transmitting valves with nickel anodes. As a very general rule, it may be stated that from one-third to one-half of the energy supplied to a valve maintained oscillatory circuit is unavoidably wasted in heat in the anode. Thus, a transmitting unit is limited in power by the heat which the valve can dissipate, and the trend of valve development, for many years, has been to produce larger and larger valves with improved devices for conveying away this wasted energy.

The power which can be safely dissipated at the anode is generally referred to as the " anode rating " of the valve.

12. **Valve Envelopes.**—The function of the envelope is, primarily, to enclose the electrodes in a space which may be evacuated, and the earliest envelopes were the glass bulbs commonly employed for electric lamps. In those valves, the heat dissipated at the anode has to be lost by radiation through the bulb, a fact which limits the maximum power which can be developed. As design progressed the glass bulbs became larger, but a practical limit was reached by the production of 1·5 kW. valves using molybdenum anodes in special heat resisting glass. Fig. 4 shows a medium powered glass valve.

Envelopes of pure fused quartz (or silica) were introduced in England for transmitting valves towards the end of the Great War. This material does not expand appreciably under change of temperature, and is, therefore, never cracked by heat ; moreover, it does not soften until its temperature reaches 1,500° C. Glass softens at about 450° C. and, as is well-known, readily cracks on sudden heating or cooling.

Silica is an ideal material for valve envelopes, and it is possible to produce silica transmitting valves of size smaller than the 1·5 kW. glass valves, yet capable of dissipating at their anodes as much as 6 kW. Many different ratings of silica valves have been standardised. Figs. 5 and 6 (*b*) show some typical silica valves of power rating from 2·5 kW. to 15 kW. ; the Service range extends upwards to 20 kW.

A particular advantage which a silica envelope has over others is that it can be opened easily for repair. The silica is cut with a carborundum wheel, and access is given to the electrode assembly; after the repair has been effected, the envelope is re-fused with the oxy-hydrogen flame and the envelope completed exactly like a new valve. Defective silica valves are thus nearly equal in value to new ones, and should be handled with just the same care.

Silica valves with anode dissipations higher than 20 kW. became possible when water-cooling of the anode was introduced. A most successful type is the one in which the anode itself is made part of the envelope. In that case the anode consists of a cylinder of copper, or chrome iron, having an open end, and a glass tube is sealed on to that end in order to complete the envelope and form an *insulating* support for the other electrodes hanging inside the anode. The anode is surrounded by a water jacket which provides the necessary cooling, as shown in Fig. 6 (a). Valves of this type have been standardised having anode ratings up to 500 kW., and large numbers of 100 kW. valves are now used in connection with broadcasting and long range W/T work.

13. **Valve Seals.**—In all types of valve, provision must be made for electrical connection to the electrodes from outside the bulb. For this purpose a conductor has to be sealed into the material of the valve envelope, a process which presents a number of special problems. One obvious difficulty consists in the difference between the coefficients of expansion of the metal conductor and the vitreous insulator forming the envelope; the difference must not be too great, or stresses will be set up which may result in the fracture of the envelope.

In small receiving valves constructed like electric lamps, all seal conductors were, formerly, platinum wires pinched in the glass stem which supports the internal parts. Platinum has practically the same coefficient of expansion as glass, and moreover, the molten glass adheres to it so that a good vacuum tight seal can easily be made. In most modern valves of this type a cheaper substitute is employed. A nickel-iron alloy has been produced with approximately the same coefficient of expansion as glass. Moreover, many new glasses have been made to withstand higher temperatures and to form seals with various standard metals. The nickel-iron alloy will not itself adhere to the glass, and so the wire is coated with copper, since the latter has been found to weld satisfactorily to glass. This **platinum substitute** or **copper-clad** wire is considerably cheaper than platinum; copper wire itself cannot be used since its coefficient of expansion is too high.

In the metal-glass valve with the water-cooled copper anode (Fig. 6), there is an additional large seal where the anode is welded to the glass portion which forms the upper half of the envelope. This big seal is made possible by means of a " Housekeeper joint " (after its inventor), which provides for compensation of the stress in such a manner that the risk of fracture is almost eliminated.

Some hard glasses have been developed having softening temperatures from about 550° C. upwards; these are specially suitable for sealing to tungsten and molybdenum wires up to 1·5 mm. in diameter. Quite large seals have been made between a chrome-iron alloy and a specially hard glass.

With silica valves, the problem of the seal is rendered more difficult by the very small coefficient of expansion of silica. Moreover, the usual glass technique of heating the vitreous material until it fuses round the wire cannot be adopted, since the softening temperature of silica is so high (1,500° C.). Actually, only one metal, lead, has been found to adhere properly to silica, and this can only be used by melting a thin pencil of lead inside a thick walled cylinder of silica. Internal and external conductors are embedded in this plug of lead at its opposite ends. Examples of these lead seals may be seen at the ends of the tubes protruding from the silica valves shown in Fig. 5. The silica appears to be strong enough, or the lead perhaps sufficiently ductile, to stand the stress due to the difference in expansion. Lead melts at 327° C., and when the conductors carry currents of the order of 50 amperes and upwards, air cooling of the seals may be required.

14. **Importance of the Vacuum. Ionisation.**—Some of the disadvantages of a " soft " valve have been referred to briefly in paragraphs 7 and 10; most of the trouble is due to the phenomenon known as " ionisation."

Everyone is now familiar with the appearance of the glow discharge tubes, widely used for advertising and decorative purposes, which are commonly (and often erroneously) called **neon tubes.** These sources of light do not contain hot filaments, but simple traces of certain gases at low pressure in the space between two electrodes in a glass tube. For more than half a century it has been known that the application of high D.C. potentials to these electrodes caused the gases to glow with colours characteristic of the particular gas. Before Fleming invented his valve, it was known that the origin of the glow is due to the impact of the electrons (from the cathode) on the molecules of gas. The electronic bombardment of the neutral molecules splits them into particles of atomic size, each associated with opposite charges of electricity, positive and negative **ions** respectively. The process is known as "ionisation by collision" and is accompanied by the emission of light, the colour of which is characteristic of the gas. In nature this effect is observed in the Aurora Borealis or "Northern Lights," an effect which is believed to be due to the ionisation of the rarefied gases of the upper atmosphere by high speed electrons emitted by the sun.

The speed of an electron depends jointly on the potential difference between the electrodes, and the mean free path between molecules of gas. If the electrons are not moving fast enough when they hit the molecules, no ionisation will occur. For this reason, if too much gas is present, the electrons cannot travel far enough between molecular collisions to acquire a velocity adequate for the produce of ionisation. If too little gas is present, electrons can travel throughout the entire space without encountering any molecules of gas to ionise

The negative ions, produced by ionisation, travel in the same direction as the original bombarding electrons, constituting, in effect, additional carriers of the electric current. The positive ions travel in the opposite direction, towards the cathode, and it is this result of ionisation which produces the most deleterious effect. When the cathode is bombarded by positive ions the active surface is usually destroyed in the case of oxide coated electrodes, or the filament is disintegrated in the case of a bright tungsten wire electrode. Moreover, the stream of negative electrons and ions flowing away from the cathode will very considerably increase the anode current, and the anode dissipation may be increased to a dangerous extent ; in addition, the currents through the valve (or discharge tube) will usually be erratic in magnitude, and uncontrollable.

In certain cases use may be found for "soft" valves (H.11), but for most purposes "hard" valves having the highest possible vacuum will be required.

15. **Evacuation Processes.**—These aim at the elimination of all occluded gases, and not only that filling the valve itself. For this reason, during the pumping process it is necessary to make all parts of the valve much hotter than they would ever become during subsequent usage. Hence, the pumping technique also involves two other main operations, a general baking process directed equally at all components, and a special heat treatment directed particularly at the metal electrodes.

While the pumps are working the valve is baked in an electric or gas oven to the highest temperatures which its materials will stand without softening of the envelope. For valves with envelopes wholly or partly of glass, the limit is about 450° C., for above this temperature the glass softens and would be deformed by the pressure of the external atmosphere. Silica valves, on the other hand, can be baked at 1,000° C. and, consequently, more thoroughly out-gassed by this baking process.

The special heat treatment of the metal parts inside the valve is achieved by electrical methods. These are of two main types—"bombardment methods" and "eddy current methods." All materials have certain quantities of various gases adsorbed on their surfaces, and absorbed or occluded inside them, and the object of all heat treatment is to cause these gases to be released.

The **E.C.H.** (eddy current heating) **process** is used for heating the metal parts of receiving valves and small transmitting valves. They are surrounded by a coil in which high frequency current flows, and the metal parts inside the valve are heated by the eddy currents thus generated in them. The coupling to the H/F coil controls the heating effect, the whole process being very rapid ; in a few seconds all the metal parts in the valve can be brought to a bright red heat. Damage to the filament is avoided by leaving the filament circuit open during the E.C.H. treatment.

The electrodes of large transmitting valves are heated by an **electronic bombardment method.** The tungsten filament is lit, and high potentials are applied to the other electrodes in turn, so that those elements are heated by electronic bombardment. The process is carefully controlled so that the gases, which are evolved in puffs, do not involve risk of ionisation, and consequent damage to the filament.

Both of the above electrode heating processes are carried out with the pumps working.

Following the evacuation process is the operation of **sealing off** the valve from the pumping equipment. A gas flame (for glass valves) or an electric arc (for silica valves) is applied to the exhausted tube quite close to the valve. This softens the tube and the external atmospheric pressure causes it to collapse inwards ; the heat is adjusted so that the tube fuses completely at the moment of collapse. A " pip " remains on the valve, and used to be visible at the top of receiving valves, and at the end of electric lamp bulbs ; in recent years, this has been concealed in the base or cap of the article.

In receiving valves, and in some low power transmitting valves, an additional process known as " gettering " is used for improving the vacuum. During assembly, a small piece of magnesium, or an alloy containing barium, is attached to a little disc of copper, or ring of nickel, which is put in such a position that it can be heated by eddy currents quite independently of the main E.C.H. process described above. **Gettering** is done either immediately before or after sealing off ; the magnesium or barium " getter " is volatilised and settles on the cold glass envelope in fine particles, forming a highly absorbent mirror which collects any minute traces of gas still remaining free in the valve, or which may be evolved subsequently. The vacuum is reduced to about one ten-millionth of an atmosphere, and is the most perfect which can be easily obtained.

Small receiving valves may be exhausted in about 10 to 30 minutes by automatic machinery. Transmitting valves, and large amplifying valves, are treated individually and carefully watched. The evacuation process usually lasts at least 5 or 6 hours ; for the most powerful valves 2 or 3 days are required.

16. **Metallised Valves.**—In many cases, when valves are used in R/F circuits, it is desirable to isolate the electrode assembly by means of an electrostatic screen. This may be accomplished by surrounding the valve with a metal conductor, such as a copper can, but is now more usually done by coating the outside of the valve with a metal composition, the latter being applied by a paint spray gun. The metallised coating may be connected to the cathode and is usually earthed, connection being taken to a separate pin in the base of the valve. With the introduction of metallised valves, receiver construction using screen grid valves (paragraph 41) was considerably simplified.

17. **Demountable Valves.**—The normal end to the life of a transmitting valve comes when the filament burns out. For many years attempts have been made to design a valve which could be opened and refitted as requisite at the transmitting station. In all such installations high vacuum pumps must be run continuously, and the station is uneconomical unless the power of the unit is high enough to justify the expense of the pumping system.

The necessity for a supply of liquid air, for use in conjunction with the mercury vapour pumps, for a long time constituted a serious difficulty. In these pumps a stream of vapour from boiling mercury is directed in such a manner that it drags molecules of air and other gases along a tube with it to a position where the mercury condenses, and the accumulated gases are removed by subsidiary pumps. The cooling of the tube connecting the mercury vapour apparatus to the valve being evacuated is usually accomplished by surrounding it with liquid air.

During 1928-29, research workers in the Metropolitan-Vickers Electrical Co., Ltd., developed oils (extracted from the various petroleum lubricants), which have exceedingly low vapour pressures at ordinary temperatures. It was found that these could be used instead of mercury in the vapour or condensation pumps, and, because of their extremely low vapour pressure, the necessity for cooling by liquid air does not arise. Demountable valves which are pumped continuously by the Metro-Vick oil pumps have been developed up to powers of 500 kW. Delay in starting up the

pumps, and the much longer delay immediately following the replacement of the filament, may be disadvantageous for some radio purposes, except in stations where there are ample facilities and space for reserve sets.

18. **Valve Failure. Signs of Age.**—Valve failures may be due to various causes, but principally they are due to failure of the vacuum, or failure of the cathode emission.

FAILURE OF THE VACUUM.—A hard valve may become soft due to a crack in the envelope, a leaky seal, or by excessive heating of the electrodes. Cracks in the glass envelope sometimes develop and are due to faulty annealing during manufacture. In the presence of much oxygen a bright tungsten filament will oxidise rapidly, producing thick yellow smoke and blue tungsten oxide. If a crack is observed in a silica valve it should on no account be lit ; the oxidation of the filament adds much to the work and cost of repair.

The first evidence of failure of the vacuum is usually a blue glow seen in the neighbourhood of the filament. The latter represents the bombardment of the filament by positive ions, a process which will ultimately lead to its disintegration. Sometimes a valve becomes temporarily soft due to the evolution of occluded gas from an overheated electrode. In those cases the gas will sometimes disappear in use, due to the " cleaning up " action of tungsten. In the latter process it appears that a little of the tungsten filament is disintegrated by the ionic bombardment and is driven on to the walls of the envelope, where it settles down and functions as a getter. The same action may sometimes occur in the case of a coated cathode, but, usually, if sufficient gas has appeared to cause a blue glow it is quite enough to ruin the cathode immediately the high potential is applied to the anode.

It will be shown later that " reverse grid current " (B.29) gives a definite indication of the state of the vacuum in the valve.

FAILURE OF CATHODE EMISSION.—Bright tungsten filaments gradually wear thin, due to volatilisation of the surface as well as to the effects of ionic bombardment. The thinning is not uniformly distributed, and is most marked at the middle of the filament where the wire is hottest ; the ends of the filament are cooled by contact with the supports and, consequently, do not volatilise so quickly. This decrease in diameter causes an increase in resistance, with a corresponding decrease in heating current and in cathode emission. If the wearing away of the wire occurred uniformly, no appreciable loss of emission would be noted, since the reduced current would be adequate to maintain the thinner wire at the original temperature. In practice, the nett effect is a sufficient reduction in temperature to reduce the total emission from the filament. This cause of loss of emission only operates after a very long life, unless the valve filament has been overheated by excessive current for considerable periods.

Occasionally a " hot spot " occurs on the filament due to some entirely local cause, such as a change in crystalline structure. This has an effect entirely different from that described above ; intense local emission and volatilisation occur, resulting in a very short life usually terminated by fusing and the formation of a brief arc.

In the case of oxide coated filaments, loss of emission is usually the result of the application of excessive potentials. Too high a voltage applied to the anode produces " stripping " of the cathode (paragraph 8).

Loss of emission is usually made apparent by an alteration in the valve constants ; g_m decreases and r_a increases (B.30).

19. **Miscellaneous Valve Defects.**—Most of the other defects which occur in valves are due to faults in manufacture. The most common defect in a receiving valve is an intermittent contact or open circuit, due to faulty welding of parts inside the valve, or faulty soldering of external leads. Some modern receiving valves, with indirectly heated cathodes, sometimes suffer from faulty insulation between heater and cathode. This is a difficult defect to locate in a multi valve receiver, since the fault is usually intermittent and may disappear on switching the supply on and off again. The evidence is usually a form of hum or rattle.

Silica valves occasionally have external leads which work loose in the lead plugs of the seals. These can be re-fixed quite easily by melting the top of the lead with a small blow lamp.

20. **The Emission Current.**—The emission from the filament (paragraph 5) varies with the heating current in accordance with a somewhat complex law. The general form of the curve is shown in Fig. 7, and is the same for each type of cathode.

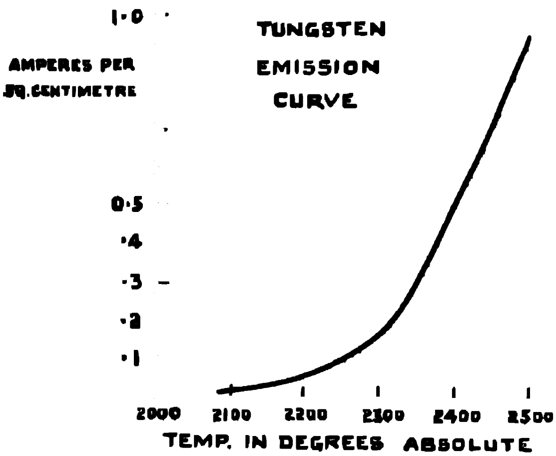


FIG. 7.

It agrees closely with the formula

$$i = AT^2 e^{-\frac{b}{T}}$$

where i = electron emission in amperes per square centimetre,
where T = temperature in degrees absolute,
where e = the base of Naperian logarithms (2·71828),
where A and b are constants which are characteristic of the cathode material.

For pure tungsten A = 60, and b = 52,400. For thoriated tungsten A = 3 and b = 30,500. The values of A and b for the "oxides" have been investigated by many experimenters, but their results differ widely. A appears to be about 0·001 and b varies from 12,000 to 20,000.

The operating temperature for pure tungsten is approximately the same as that of a filament in an electric lamp. Thoriated tungsten valves (paragraph 7) have filaments distinctly less bright, though definitely glowing, and hence arose the name "**dull emitter**" when this type of filament was introduced. The same term is used for the oxide coated filament valves, and is more justified in this case, for the operating temperatures are so low that in certain types of valve the glow from the filament is practically nil.

The steepness of the curve indicates the rapid increase in temperature of the filament with heating current, and is an indication of the cause of the shortness of life to be expected in a valve if the filament temperature is not carefully controlled.

21. **The Anode Current.**—The anode current, measured by the milliammeter in Fig. 1 (a), will not be equal in value to the emission current unless the anode potential is high enough to attract every electron emitted by the filament. Under normal conditions, the anode current increases as the anode voltage is raised up to the potential at which the anode attracts all of the emitted electrons. After that point no increase in anode current is possible, no matter how much further the anode voltage is increased.

The limiting value of anode current is called the **saturation current,** and is the total available filament emission. The anode current can only be increased, for higher anode voltages, by increasing the filament temperature and thus increasing the available emission current.

The rise in anode current, with *fixed* filament temperature and increasing anode voltage, follows curve of the form shown in Fig. 8, and is found to follow the law

$$I = \frac{K}{r} . V_a^{\frac{3}{2}}$$

here K = a numerical factor,
here r = radius of the cylindrical anode surrounding the filament,
here I_a = anode current,
here V_a = anode voltage.

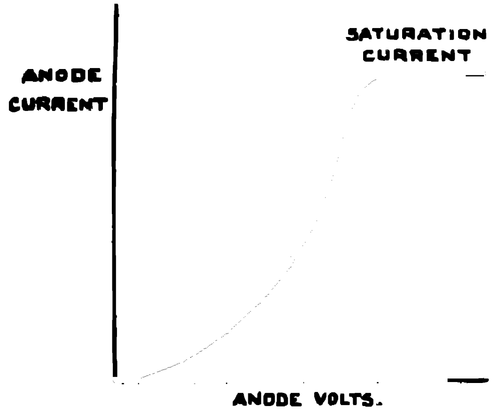


FIG. 8.

This is an important relationship and is known as the " **three-halves power law.**" The curve rises until the anode current attains the saturation value, when it bends over and becomes horizontal, except in the case of coated cathodes which do not show a steady saturation value, the emission current continuing to increase with increase in anode potential, but not in accordance with the formula quoted above.

The saturation current is never reached in any serviceable thermionic valve, as the filament is designed to give ample emission, easily covering the highest anode voltage which may be applied. In order to observe the saturation effect in a standard type of valve, it is necessary to reduce the filament current and so limit the available emission.

In a valve with more than two electrodes, the emission current can be obtained by connecting the terminals of all the other electrodes—except, of course, the cathode— to the anode terminal, and thus treating them as one collecting electrode. When the cathode is a coated one, care must be taken that it is not stripped by the application of too high a voltage at the adjacent electrode.

The behaviour of the anode current may be investigated from another point of view. Let us consider the voltage between anode and filament as constant, and raise the filament heating current and hence the filament temperature. For low values of filament temperature the available emission is so small that the applied anode voltage will attract all the electrons ; these conditions will hold with increasing filament temperature, until the latter rises to such a point that the emission current just equals the current the anode is capable of attracting. This effect is shown in Fig 9, which represents the maximum anode currents which the respective anode potentials are able to collect with varying filament temperatures. As the temperature is increased still further, only a portion of the emitted electrons are collected by the anode and the remainder fall back towards the filament.

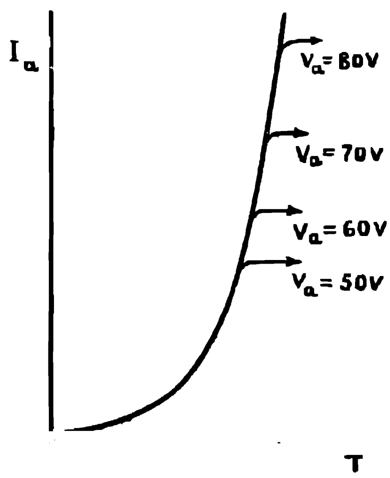


FIG. 9.

Emitted electrons which are in excess of the demands for anode current remain in the neighbourhood of the filament, forming a negatively charged cloud which is known as the " **space charge.**" Further heating of the filament simply makes the space charge become more dense, the anode current remaining at a steady value. The presence of a space charge has the effect of preventing some of the emitted electrons from coming under the influence of the attractive force of the anode ; partial neutralisation of the effect of the space charge may sometimes be effected if positive ions are present (H.11).

22. **The Diode, Uses of Hard and Soft Diodes.**—The form of the characteristic curve of a diode, showing the relation between the anode voltage and the electron current flowing from filament to anode, is the same as that of the simple curve shown in Fig. 8, and is described by the

three-halves power law. The diode is the simplest of the valves, and has a number of uses which include :—

(*a*) Power rectification of A.C. supplies (H.2-3).
(*b*) Detection of modulated wave forms in high quality R/T receivers (N.27).
(*c*) The provision of the various forms of A.V.C. (N.35).

X-ray tubes are also diodes, but discussion of their design and characteristics is outside the scope of this work.

An interesting static characteristic of a high power silica rectifier is shown in Fig. 10. The curve shows signs of reaching saturation at about 2·5 amperes emission, with about 1,500 volts on the anode, but instead of becoming horizontal it continues to rise. This is because the anode is becoming hot by the bombardment of the electron current, and, consequently, the loss of heat from the filament is restricted ; the filament becomes hotter and it emits more electrons.

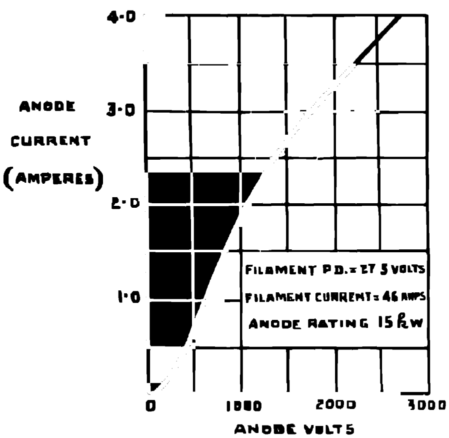


FIG. 10.

This type of valve is used in high power rectifying panels (H.4), and would supply power with a loss of about 1,000 volts in the diode. The mean current would then be of the order of 2 amperes, and the A.C. resistance of the valve approximately 500 ohms. The A.C. resistance is made as small as possible by making large both the emitting surface of the cathode and the collecting surface of the anode, and locating each as close to the other as possible.

In simple full wave rectifiers, such as those used to give the requisite potentials in a receiver, the two diode valves are usually included in one bulb ; these **double diodes** produce economy both in space and in cost, and usually employ the same cathode.

Hard diode power rectifiers involve considerable loss of power ; for this reason, **soft diode** mercury vapour rectifiers have been coming into favour in recent years (H.11). The deleterious effect of ionic bombardment of the cathode is countered by suitable control of the potential drop across the valve.

23. **Introduction of the Grid. The Triode.**—The third electrode introduced by de Forest in 1907, and named by him the " grid," consists of an open wirework, either a spiral or mesh, and is placed between the cathode and the anode. The potential of the grid relative to the cathode exercises a controlling effect on the electron current of the anode, while, due to its open construction, it does not act to an appreciable extent as a collecting electrode. In practice, the potential of the grid is varied ; incoming signals produce voltage variations across the tuned circuits which are ultimately applied between grid and cathode of a valve, with consequent variations of the anode current at the same frequency. Until about 1924 the only type of valve in common use was a triode with a filamentary cathode.

The very important position which the three-electrode valve occupies in wireless engineering, and in allied sciences, is due to the remarkable control which the grid exerts on the anode current. This is illustrated by the graph, Fig. 11, which shows how the anode current of a standard type of receiving valve varies with changes in grid voltage.

Starting with the grid potential at a considerable negative value with respect to the negative terminal of the filament, and gradually reducing this negative value, it is observed that no anode current will flow, although the anode potential is 50 volts, until the grid potential reaches —5 volts. As the grid potential changes from —5 volts to zero the anode current rises to about 0·7 mA., and

it continues to rise as the grid voltage increases in the positive direction. The anode current, in the example shown, attains the saturation value, of approximately 1·5 mA., when the grid voltage reaches 6 volts positive. This curve describes the relationship, in this particular valve, between the anode current and the controlling grid voltage. It is usually termed a MUTUAL CHARACTERISTIC of the valve, and a family of such curves taken at various anode voltages can be used for ascertaining the behaviour of the valve under various circuit conditions.

The grid itself begins to take current at about zero grid voltage, but this current is *very* small at first. In the case shown in Fig. 11, for example, the grid current is only about 100 microamperes when the anode current has attained its maximum value of 1·5 mA. at about 5 to 6 volts. Thus it would appear that, for negative and small positive values of grid voltage, the grid exercises considerable control over the anode current with practically no expenditure of energy itself, *i.e.*, with a minimum of damping in the input circuit to the grid. But this question is complicated by the inter-electrode capacities, owing to which the output circuit may be responsible for power being consumed by the input circuit, even when no grid current is flowing. However, with properly-designed apparatus this effect can be made a minimum, and the fullest advantage taken of the amplifying property of the triode. This is the property of the thermionic valve with a control grid which gives it its unique position as an amplifier in wireless reception.

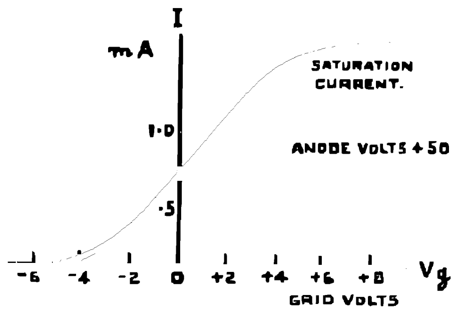


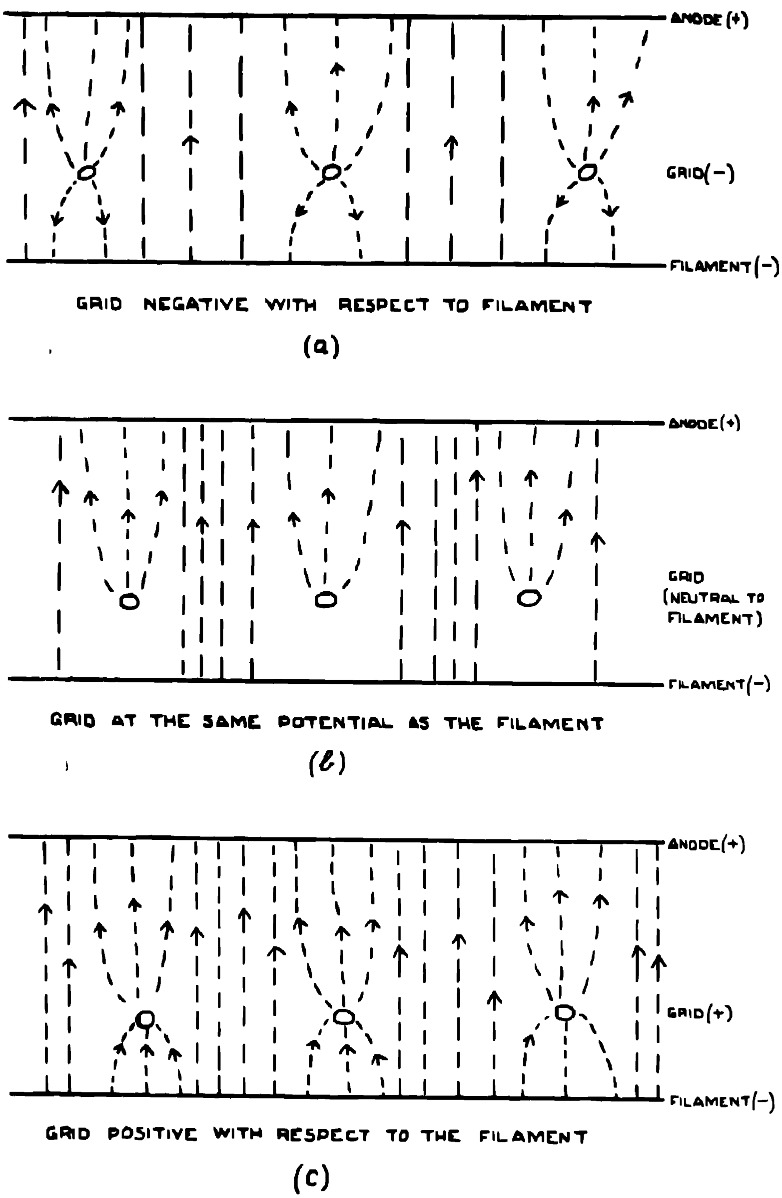
FIG 11.

A mental picture of the action of the control grid may be obtained with the conception of lines of force as illustrated in Fig. 12, which shows enlarged diagrams of portions of the three electrodes. These lines represent the idea of strain, or attraction between positive and negative charges at the ends of the lines. Thus lines of force connect positive charges on the anode with negative ones of the filament or in its vicinity. In Fig. 12 the electrostatic field is shown between the electrodes without the presence of electrons, in order to make the position more easily understood. The arrowheads indicate the direction in which electrons tend to move, and not the direction of conventional current.

When the grid is negative with respect to the filament [Fig. 12 (*a*)], the lines of force between the grid and filament represent the strain in the reverse direction, *i.e.*, electrons are forced back to the filament by the negative grid potential. If the grid potential is sufficiently negative, the influence of each wire will extend along the greater proportion of the filament and entirely suppress the emission current.

When the grid potential is neutral with respect to the filament there is no electrostatic strain between the two electrodes, and electrons are therefore removed solely by the strain represented in Fig. 12 (*b*) by the lines of force joining anode and filament between the grid wires.

When the grid potential becomes positive, Fig. 12 (*c*), the direction of the field between grid and filament is everywhere such as to draw electrons away from the filament, but, because of the large spaces between the grid wires, the majority of the lines of strain connecting filament and grid are curved. All the electrons which are initially given an outward impetus by the grid potential travel faster and faster under this attraction, and because of their acquired momentum the majority of them do not keep to the original curved lines of strain, but are shot through the open spaces between the grid wires. Practically, only the electrons which are drawn straight to the grid wires actually reach them, and hence the grid current is small. All the others pass through the interstices of the grid and travel on to the anode. The greater the grid potential, the greater becomes the field intensity near the filament, and the quantity of electrons removed to the space outside the grid is correspondingly increased.



ELECTROSTATIC FIELD IN A TRIODE

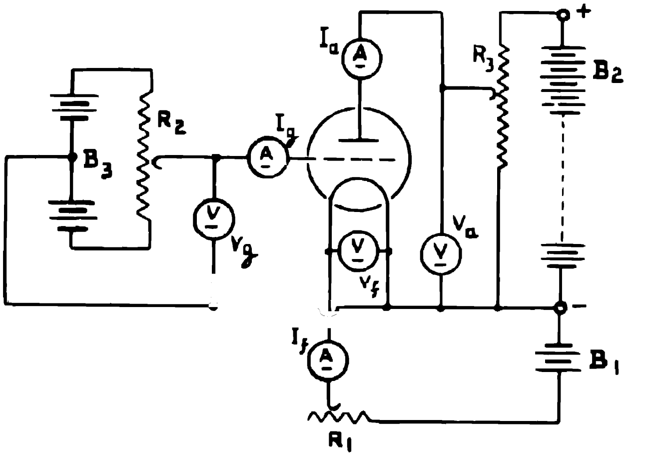
FIG. 12.

At high positive values of grid potential—approaching the anode potential—the concentration of the field considerably increases the grid current, and if the grid potential exceeds the anode potential there will be a reverse field outside the grid tending to send back to the grid the electrons which have passed through. The grid current will then be very high. Such an effect does not occur in valves operating in receiving apparatus, but conditions approaching it sometimes occur in transmitting valves (K.61).

24. **Secondary Emission.**—It has been shown by numerous experimenters that when electrons hit an electrode, such as the grid or anode of a valve, they may cause the ejection of other electrons by their impact. These "**secondary electrons**" may actually exceed in quantity the primary electrons which produce them (K.11). Obviously they can only be detected by being collected on an appropriate electrode at a higher potential than the target electrode. Secondary electrons emitted by the anode of a triode at high potential are attracted back to the anode by this potential, but secondary electrons emitted by a grid, which happens to be at a potential sufficiently high to cause the production of secondary emission, may pass to the anode which is normally at a still higher potential, and thus the secondary electrons are included in the anode current. The effect, in general, is very undesirable, and numerous efforts have been made to prevent it by coating the grid with various materials, but no satisfactory solution is known at present. A particular effect which sometimes arises with a transmitting valve, and is known as "**blocking,**" is an example of limitation of operation by secondary emission. The conditions of a transmitting circuit cause

the grid potential to swing positive and negative. If these conditions produce such a relatively high positive swing that the grid current becomes very large, the resulting secondary emission from the grid may make the grid potential so much more positive that it cannot return. The result is a cessation of oscillations in the circuit and a great increase in both anode and grid currents at high static potentials, with consequent great risk of overheating one or both electrodes (K.11).

25. **Triode Characteristics. Valve Testing Circuit.**—It has been shown in the preceding paragraphs that there are two currents flowing in a triode, the anode current and the grid current. These vary with the cathode emission, the grid voltage, and the anode voltage. It is usual to represent the relationships between these five variable but mutually independent quantities by graphs, some of which are of great value in estimating the behaviour of the valve under working conditions. These graphs are termed "characteristics" of the valve. Each graph is one showing current variation, plotted against voltage variation, and, obviously, several types of such curves may be produced, and it is necessary to select those of most value for the particular purpose.



VALVE TESTING CIRCUIT.

FIG. 13.

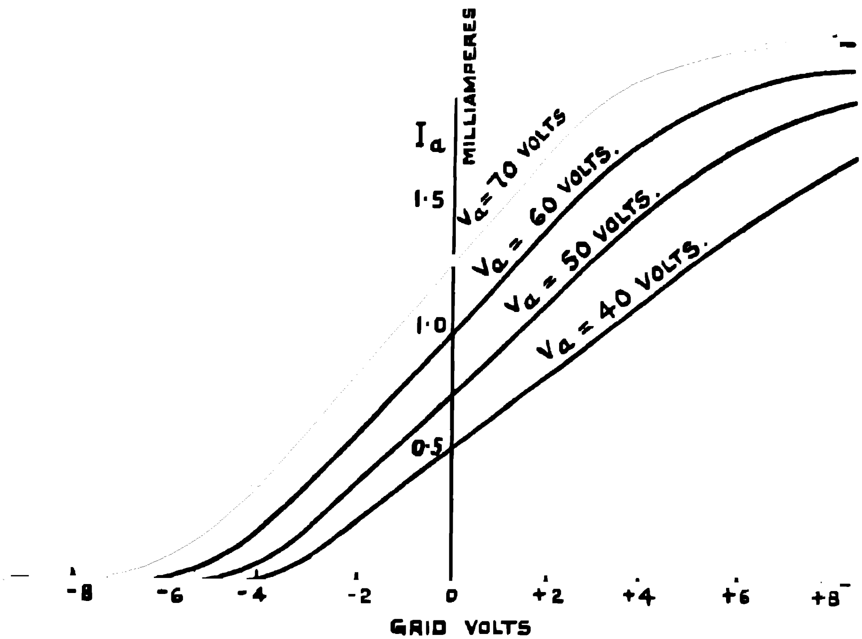
Fig. 13 shows a simple circuit suitable for testing directly heated triodes. The figure is self explanatory, and indicates that variable voltages may be applied to the filament, grid, and anode, and the corresponding currents observed in suitable ammeters. It will be noted that all voltages are measured with respect to the L.T. negative terminal, and that ammeters are placed in such positions that they do not carry, and give inaccurate results by indicating, the currents flowing through the corresponding voltmeters. This is particularly important in the grid circuit, where the current between filament and grid through the valve may be only a few microamperes.

Valve test boards designed to test all kinds of multi-electrode and multi-purpose valves will vary in complexity with the number of tricks they can perform. Considering any given use of such a board, there will always be an essential simplicity illustrating the above principles, and no useful purpose would be served by considering the diagram of a comprehensive test board.

In practice, the valve manufacturer always specifies the optimum conditions under which the valve should be operated. It is usually of importance to adhere to these conditions, particularly in regard to those which determine the heating of the filament.

26. **Mutual Characteristics.**—The curves of most general use for triodes are those showing the effects on the anode current due to changes in the grid voltage, referred to in paragraph 23; these curves are termed the **mutual characteristics** of the valve, and a typical set for a small triode used for reception purposes is given in Fig. 14.

In determining a family of these curves it will usually be most convenient to set the grid bias at a fixed value, and to observe the anode current given by different anode voltages. This determines a number of points, one on each curve, and further points are obtained by setting the grid bias to some other value and obtaining a second set of readings of



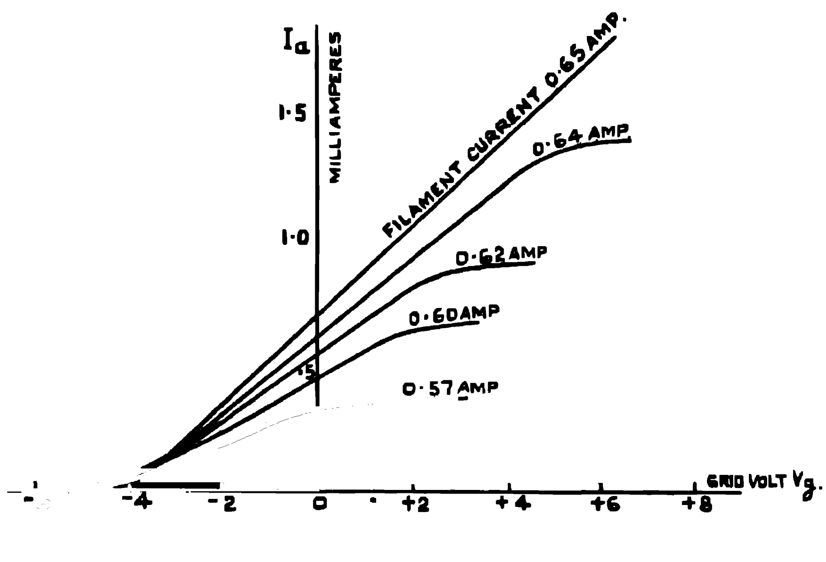
Mutual Characteristics of Triode.

FIG. 14.

anode current. In this way a complete set of readings may be obtained, and the curves drawn from them.

The lower portion of the characteristic follows, approximately, the " three-halves power law " (paragraph 21), but the graph becomes a straight line and continues as such until the anode current approaches the value of the maximum available emission, or saturation current, when it curves over and becomes almost horizontal. As a general rule it also falls as the positive grid potential is increased, because the grid itself collects electrons and so reduces the current flowing to the anode (*cf.* paragraph 29).

The graphs corresponding to higher anode potentials are of a form similar to those taken at lower potentials, but displaced towards the left of the diagram, as shown in Fig. 14 The straight portions are nearly always of increasing steepness and attain very slightly higher saturation values as the anode voltage is increased.



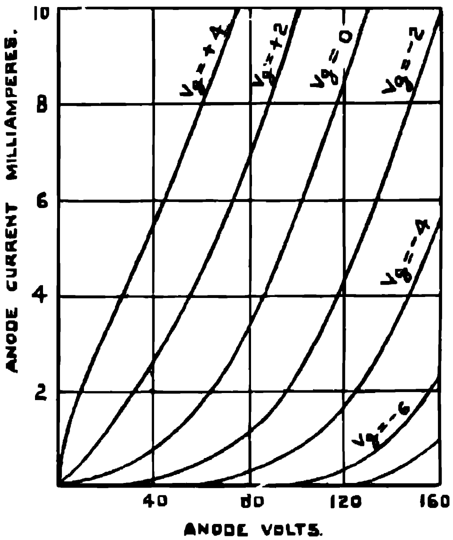
ANODE CURRENT - GRID VOLTS CHARACTERISTIC CURVE (VARYING I_f).

FIG. 15.

If the filament current is increased, the maximum value of the anode current is raised and, in addition, the slope of the straight portion of the characteristic is slightly increased — Fig. 15. These two effects, viz., increase of maximum current and increase in " slope," either together or separately, sometimes have the effect of improving the operation of a receiver or transmitter as will be understood from subsequent discussion. There is, therefore, a

temptation to overrun filaments to obtain better results. Occasions may arise when this technical offence may be justified by the necessity for achieving certain results, but the risks which have been previously mentioned in this section will have to be remembered.

27. **Anode Characteristics. The Load Line.**—A family of curves called the **anode characteristics** are obtained by plotting anode current against anode voltage, the grid voltage being kept constant for each curve. Fig. 16 shows a typical set of these characteristics for a small receiving valve.



ANODE CHARACTERISTICS OF TRIODE.

FIG. 16.

The filament of the valve, in the example shown, provides ample emission, and therefore no saturation is reached on these graphs, which indicate at least 10 milliamperes of available emission. If the filament had (say) 8 milliamperes of available emission, the curves would bend over and become horizontal at that value; they would, in fact, be special cases of the general form shown in Fig. 8. Moreover, if the valve were a transmitting valve, there would be some very evident limiting value of positive grid bias on exceeding which little further increase in anode current results. Under those conditions there would be a characteristic curve on the left of the graph which is usually known as the "**limiting edge.**"

The anode characteristics are, in general, probably more useful than the mutual characteristics for determining the behaviour of a valve under various circuit conditions; valve manufacturers are now including them in their catalogues more than formerly.

The anode curves can easily be drawn from the mutual curves. Thus, in Fig. 14, a group of readings may be taken where the characteristics cut the vertical ordinates at (say), $V_g = +2$, $V_g = 0$, $V_g = -2$, etc., and each group gives the data for one of the curves in a family such as Fig. 16.

As before, when taking characteristic curves of valves in the laboratory, it is more convenient to take the series of readings with fixed grid volts and to vary the anode volts; any other method is troublesome, since changes in grid potential produce anode current alterations, and the alterations of potential in the anode potentiometer produce changes in the anode voltage.

As an example of the use of anode characteristics, consider the simple circuit of Fig. 17 (*a*), and the corresponding graphs, Fig. 17 (*b*). The triode has a load equivalent to a resistance R in the anode circuit, and power is supplied from a high tension battery of voltage V. Let V_g represent the voltage applied to the grid.

If V_g is adjusted to a steady negative value so that I_a falls to zero, the voltage on the anode will be V. Let A [Fig. 17 (*b*)] represent these conditions.

If a current I_a is allowed to flow in the anode circuit by adjusting the grid bias, the voltage drop between the high tension supply and the anode will be $I_a \times R$, and the anode voltage becomes $V - RI_a$. Suppose the point C on the graph represents the new conditions. Draw the straight line ABCDEF, cutting the characteristics at the points indicated by these letters. Draw CH perpendicular to OA. OH obviously represents the voltage on the anode which is $V - RI_a$, and CH represents the corresponding current I_a; AH represents the voltage drop in the external load, *i.e.*, $AH = RI_a$.

$$\therefore \quad R = \frac{RI_a}{I_a} = \frac{AH}{CH} = \cot \theta.$$

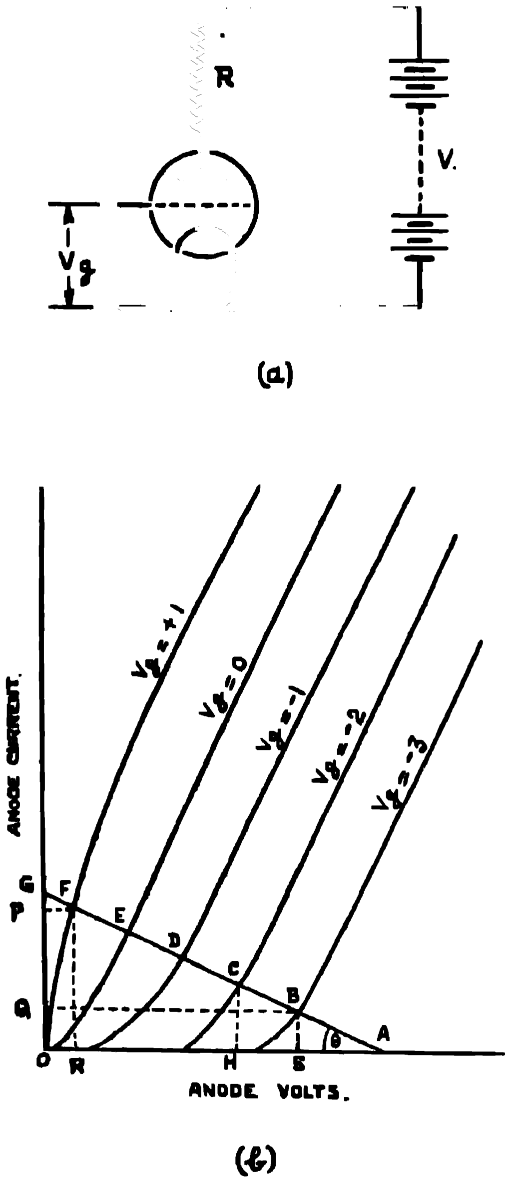


FIG. 17.

R is constant, *i.e.*, cot θ is constant, and therefore the series of conditions brought about by a swing of grid voltage of amplitude AF, will be represented by the straight line ABCDEF, called a "**load line.**" It is a characteristic of the circuit and not of the valve alone. The load line is drawn across the anode characteristics at an angle representing the external conditions, and is used to give information concerning the amplitude of grid swing and the external conditions of anode current and voltage.

When a valve is receiving an *oscillatory input*, the mean bias potential of the grid with respect to the filament is either determined automatically, or else held at a pre-determined value. Thus, suppose the **working point** is determined by the mean grid bias of —1 volt, it may be represented by the point D on the graph ; the load line is drawn through D and the angle θ determined by cot θ = R. As the grid swings about the working point D, corresponding changes will take place in the anode current. It does not, however, follow that *equal* changes in grid voltage will be accompanied by *equal* alterations of anode current ; this will only be the case if the load line made equal intercepts FE, DE, etc., with the various characteristics. It is these inequalities of anode current variations which produce **distortion** when a valve is being used as an amplifier.

The **undistorted A/F power output** of a broadcast receiver is a matter of some importance, and one to which the manufacturer always draws attention. It may be easily measured by an " output meter " or estimated from the load line applying to the output stage of the set, providing that one is given the magnitude of the oscillatory input between grid and filament. The mean power output may be determined from the peak values of the anode current and anode voltage swing. Thus, in Fig. 17 (*b*), the anode current varies from P to Q, the anode volts similarly varying from *point* R to S. If the amplitude of the anode current is given by $\mathcal{I}_a$ and that of the anode volts is given by $\mathcal{V}_a$, we have $\mathcal{I}_a = PQ/2$ and $\mathcal{V}_a = RS/2$. The power output is given by the product of the R.M.S. current and the R.M.S. volts, hence,

$$\text{Power output} = \frac{\mathcal{I}_a}{\sqrt{2}} \times \frac{\mathcal{V}_a}{\sqrt{2}}$$

$$= \frac{PQ \times RS}{8} \quad \ldots\ldots\ldots$$ in watts, if $\mathcal{I}_a$ and $\mathcal{V}_a$ are in amps. and volts respectively (*cf.* N.47).

When a valve is used as an A/F power amplifier, the load line is adjusted to give intercepts as nearly equal as possible with equally spaced anode characteristics. The valve manufacturer usually specifies the optimum impedance to be used with any output valve (N.46).

28. Numerical Example on the Load Line.—The anode characteristics of a triode output valve are given by the following practical readings, and represented graphically by Fig. 18.

At $V_g =$ 0	V_a	20	40	60	80	100	volts
	I_a	2	6·5	14	24	33	mA.
At $V_g = -$ 14	V_a	100	120	140	160		volts
	I_a	1	5	11·5	19·5		mA.
At $V_g = -$ 28	V_a	180	200	220			volts
	I_a	1	4	9			mA.

The valve is used with an H.T. battery of 200 volts, a grid bias of — 14 volts, and a static load resistance of 4,500 ohms. Draw the load line and calculate the power output in milliwatts if the input signal has a peak voltage of 14 volts.

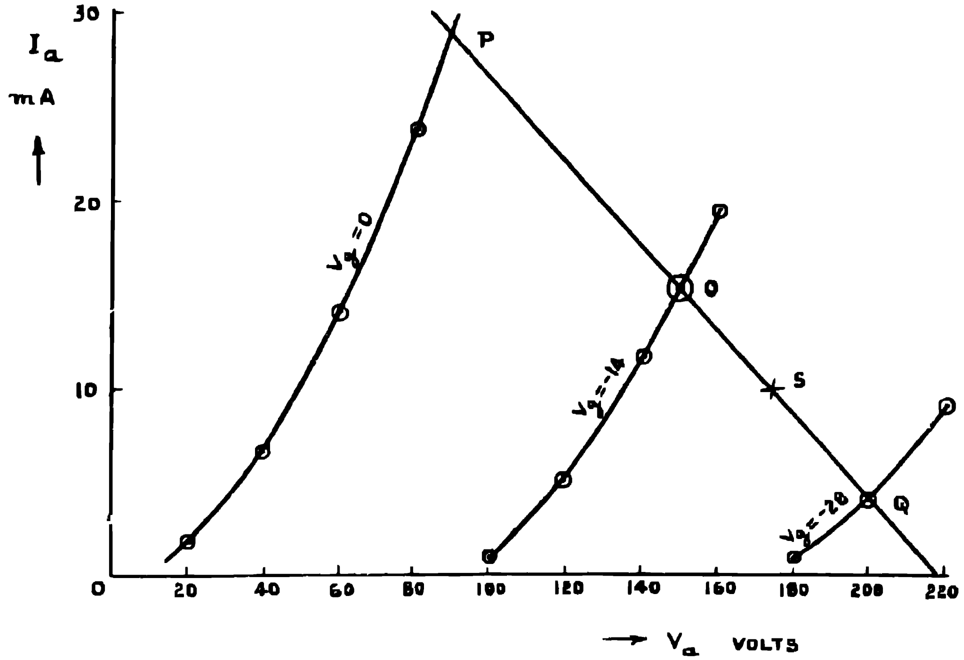


FIG. 18.

When the anode current (I_a) is zero, the volts on the anode (V_a) will be 220 volts. When I_a is 10 mA., the voltage drop in the load resistance will be given by $10 \times 4{,}500 \times 10^{-3} = 45$ volts. In that condition we have $V_a = 220 - 45$ volts $= 175$ volts.

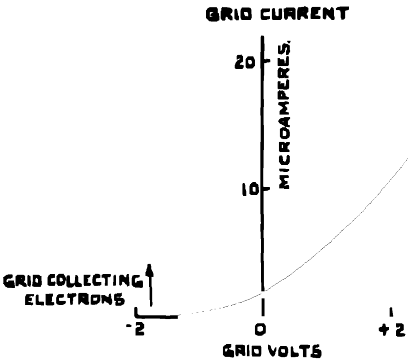
The above conditions determine the load line POQ which starts on the X-axis at 220 volts, and passes through the point S determined above. The load line intercepts the anode characteristic for $V_g = -14$ at the point O which is thus the " **working point.**"

The grid swings with an amplitude of 14 volts about the working point. This gives an anode voltage swing of $200 - 90 = 110$ volts, and an anode current swing of $29 - 4{\cdot}5 = 24{\cdot}5$ milliamps

This gives .... **power output**$= \dfrac{110 \times 24{\cdot}5}{8} = 337$ **milliwatts.**

29. **Grid Characteristics.**—The grid functions primarily as a controlling and not as a collecting electrode, yet, in certain cases, the flow of grid current is essential.

When the grid potential is more than a fraction of a volt negative, electrons are not collected. The electron current begins in the vicinity of zero grid voltage, the actual point of commencement being dependent upon the nature of the materials constituting grid and filament. Owing to the phenomenon known as **contact difference of potential,** there is a P.D. between the surfaces of the grid and filament whenever those two are in contact and made of different materials; they are in metallic connection through the grid circuit. Thus, when the filament is tungsten and the grid is nickel or molybdenum, there is a trace of grid current even when the applied potential on the grid is slightly negative—Fig. 19. With a coated filament, however, the grid current does not start until the grid potential is half a volt or so positive. The value of the current is very small—a few microamperes only—for low grid potentials, because of the comparatively small dimensions of the grid wires.



HARD VALVE

GRID CURRENT CURVE OF TRIODE.

FIG. 19.

It is common practice to show the grid characteristics on the chart of mutual characteristics, in which case a magnified scale must be used for the grid current.

As the grid voltage is made more and more positive the anode current increases, and when the voltage exceeds that on the anode the grid current may be very much in excess of the anode current. Fig. 20 (*b*) shows the curves of grid current for fixed anode voltages and varying grid volts. From these it is evident that the anode voltage has little effect on the grid current except at low values of anode volts. If the grid voltage is increased very considerably with respect to the anode volts, the graphs tend to curve *downwards*; this phenomenon indicates the beginning of the secondary emission which leads to **blocking** in transmitters (K.11). It is not usual to draw grid characteristics beyond a few positive volts.

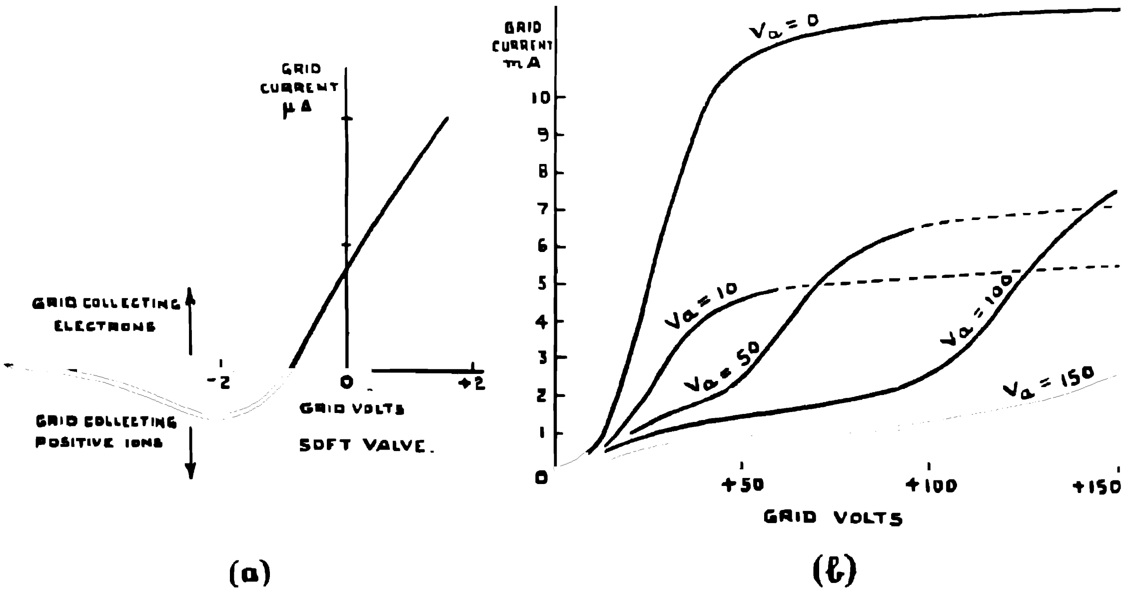


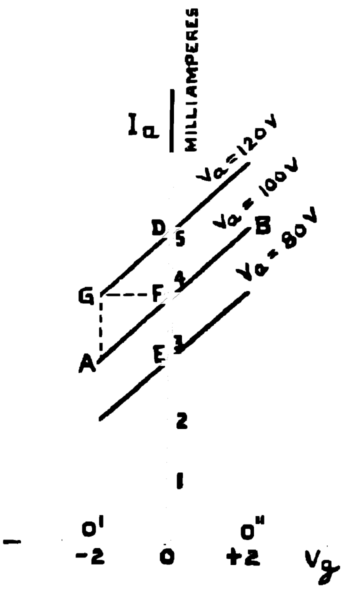
FIG. 20.

The grid characteristic is a very delicate indicator of the state of the vacuum inside the valve. If any ionisation occurs, the positive ions travel away from the anode towards the grid and filament. If the grid is at a negative potential with respect to the filament, it attracts these ions and collects their positive charges. The resulting current is in a direction opposite to that due to collection of electrons, and is therefore termed "**the reverse grid current.**" The term "**backlash**" has also been used for this reverse current, but it is gradually being replaced by the former expression.

Fig. 20 (*a*) shows the grid characteristic of a slightly soft valve. It is standard practice to specify the maximum value of the reverse grid current at a fixed value of anode current and/or corresponding anode voltage. Thus receiving valves, in general, are tested at -2 grid volts with 100 volts on the anode, when the reverse grid current must not, for example, exceed 2 microamperes. In the case of a certain high-power transmitting valve, the corresponding limit is 50 microamperes at -20 grid volts, with 1 ampere of anode current at about 7,000 volts.

The reverse grid current falls to zero as the anode current is reduced by an increased negative potential on the grid. This is shown in Fig. 20 (*a*) for a small receiving valve, where the current is reduced to zero by only a few negative volts. In the case of a large transmitting valve, a negative grid bias of several hundred volts may be required to suppress the current.

30. **The Triode Constants.**—As has been explained in the earlier portions of this section, the outstanding property of the triode is the control of the anode current by small changes of grid voltage. This control is most fully expressed by means of the various characteristics. It is customary, and useful, to deduce certain particular pieces of information from those curves in order to express briefly the relationship between changes in the grid and anode voltages, and the anode current, under stated conditions. These **valve constants** are usually deduced from the straight portions of the mutual characteristics, but may be referred to other points also. The straight portions are very nearly parallel, and from their geometry the required valve constants may easily be obtained, although it is necessary to specify the conditions of anode and grid voltage at which the observations are made. For receiving valves these are usually $V_a = 100$ volts and $V_g = 0$.



MIDDLE PORTIONS OF MUTUAL CHARACTERISTICS OF A TRIODE

FIG. 21.

31. **Mutual Conductance** (g_m).—The most important constant is that which describes the rate of change of anode current with change of grid voltage, the anode voltage remaining constant. It is termed the "mutual conductance," and the usual Service symbol is g_m. It is the gradient or slope of the mutual characteristic at a particular point, and is usually expressed in milliamperes per volt; for this reason g_m is sometimes briefly referred to as the "**slope of the valve.**"

Fig. 21 represents the middle portions of the mutual characteristics of a triode. The graph AB rises from 3 mA. to 5 mA. as the grid voltage changes from -2 volts to $+2$ volts. The value of g_m is therefore $0 \cdot 5$ mA. per volt. It is to be noted that *changes* in current and voltage are used in obtaining the constant, *not the actual values* of current and voltage.

Although the mutual conductance represents the control of the grid voltage over the anode current, for consistency it is measured under the *static* condition of constant anode voltage. When there is an external

impedance in the anode circuit, the voltage drop across the impedance alters with changes in anode current, producing corresponding changes in the voltage between anode and filament. Consequently, a given change of grid voltage cannot produce in such a circuit the full change in anode current which is indicated by $g_m \times V_g$, where V_g denotes the *change* in grid volts.

A large number of valves in receiving apparatus operate at, or near, zero grid voltage, and are designed to have characteristics showing straight portions at this value of grid potential. The mutual conductance of such valves is almost invariably measured at $V_a = 100$ volts and $V_g = 0$. Many valves, however, are designed to operate at a mean negative grid voltage, often of high value. In such cases the straight portions of the mutual characteristics in use are often entirely to the left of the zero grid volt line ; the mutual conductance is then usually referred to the mean grid bias and the maximum anode voltage employed with the valve.

The point at which a mutual characteristic intersects the axis of grid volts is usually referred to as the " **cut-off** " point ; it is the point at which the anode current is reduced to zero, and the negative grid bias which is required to achieve this condition is called the **cut-off bias.**

32. **Variable Mu Valves.**—It is possible to modify the design of the triode so that the grid has less complete control over the flow of electrons between cathode and anode. If a gap is left in the grid winding, or if the cathode is allowed to protrude outside the ends of the grid coil, the mutual characteristic alters in form and acquires a long " tail." High values of negative grid bias are then required in order entirely to suppress the anode current, and the general feature is that a sharp **lower bend** and cut-off point becomes replaced by a smooth curve having a gradual change of mutual conductance from high to low values. In many cases the cut-off bias in a valve used for receiving purposes may be from 20 to 40 volts negative.

Valves with these characteristics are called " variable mu valves " ; in this country the name is taken to refer to the smooth change which takes place in the mutual conductance as the bias point is gradually altered. It will be seen later that this smooth change can be used to good advantage in controlling the **stage gain** of an amplifier. Triodes may have variable mu characteristics, but they are usually only used with tetrodes or pentodes.

Typical variable mu characteristics are shown in Fig. 28 of Section " N." In one commercial screen grid R/F amplifying valve, the value of g_m is quoted as varying from 2·5 to 0·03 mA. per volt, under certain stated conditions.

33. **A.C. Resistance or Impedance** (r_a).—The anode current may be changed by changing the anode voltage, while keeping the grid voltage constant. In the early days of the use of triodes the " **anode conductance** " was therefore deduced by observations on change in anode current at zero grid volts. But it is more convenient to use the reciprocal of this when considering the behaviour of the valve and its associated circuits, for it is usually a change of anode current which causes a change of anode voltage, and not the reverse. The term " Impedance " has been widely used for many years for this constant. However, many authorities have pointed out that the term *impedance* includes the effects of inductance and capacity, and is dependent on frequency, while the constant of the valve to which the term has been applied excludes inductance and capacity, and is always determined from steady current readings, so that it does not depend on frequency. Several terms, including the word " resistance," have been proposed, and it appears that " A.C. Resistance " (anode characteristic resistance) is likely to meet with the most general agreement. The symbol is r_a.

Referring again to Fig. 21, the anode current at zero grid volts rises from 3 mA. at 80 anode volts to 5 mA. at 120 anode volts. The A.C. resistance or impedance at 100 volts is therefore 40 volts/2 mA. = 20,000 ohms.

The A.C. resistance may be defined as the reciprocal of the rate of change of anode current with anode voltage, when the grid voltage is kept constant. Again, *changes* of anode voltage and current are referred to, *not the actual values*.

34. **Amplification Factor (*m* or sometimes μ).**—A change in anode current brought about by a change in grid voltage is much greater than the change in anode current brought about by the same change in anode voltage. The ratio of the two is called the "amplification factor," the usual Service symbol being m.

The practical derivation of this constant may be simply illustrated by means of Fig. 21. Suppose the conditions are defined by the point F on the 100 volt characteristic. Let the grid voltage be changed to -2 volts while the anode voltage remains steady. The anode current falls to the point A. Now raise the anode voltage, keeping the grid potential at -2 volts, until the current attains the value it had under the first conditions. Let G be the point representing this final condition. It is seen from the characteristics that a change of 20 anode volts is required to counterbalance the effect of a change of two volts on the grid. In other words, the grid control is 10 times as effective as the anode control, or the amplification factor is 10.

The amplification factor is the ratio of two similar units and is therefore a pure number, for which no units can be specified.

35. **Relation between g_m, r_a and m.**—From the above discussion, the simple relationship between the three constants now becomes apparent. By definition, the amplification factor is the ratio between the mutual conductance g_m, and the anode conductance. Now the anode conductance is the reciprocal of the A.C. resistance r_a, hence $m = g_m \times r_a$. This result may also be deduced geometrically from Fig. 21.

The mutual characteristics of a simple triode become slightly steeper with increasing anode voltage; in other words, the mutual conductance increases. In that case it follows that characteristics representing equal increments of anode voltage cut the zero grid volts ordinate in increasing lengths of sections, *i.e.*, the anode conductance increases, or the A.C. resistance decreases, with increasing anode voltage. When the mutual characteristics are close together, the A.C. resistance will be high.

Hence, the product of g_m and r_a tends to remain constant over a wide range of anode voltage, but falls slightly at very low values, particularly when the grid is highly negative at the same time. Over a large portion of the characteristics the variation of amplification factor is not greater than about 10 per cent.

36. **Valve Constants and Electrode Design.**—Valve constants are determined by the geometry of the electrodes, and fairly accurate design formulæ have been developed by various authorities. The calculations are beyond the scope of this work, but the general principles may be appreciated.

The amplification factor is high if the mesh of the grid is close (*i.e.*, the wires close together), and the distance from the filament (or cathode) to the anode is great compared with the distance to the grid. The mutual conductance increases with the value of m, but is nearly inversely proportional to the distance between the emitting surface and the grid. The A.C. resistance increases with increasing values of m, and also increases with increasing distance between the emitting surface and grid. The greater the proportion of anode surface actually collecting electrons, the greater is the mutual conductance and the smaller the A.C. resistance (paragraph 22).

Valves are designed to fulfil particular functions in wireless apparatus, and the constants may be used to determine their most suitable functions. Thus a valve with high values of r_a and g_m could be employed as an anode bend detector, or an R/F amplifier. Similarly, a low impedance valve with a good value of mutual conductance is desirable in the last or output stage of a receiver.

The following table shows the ranges of values of the constants of typical triodes :—

(i) FILAMENT VALVES.

Purpose.	A.C. Resistance (Ohms.)	Amplification Factor.	Mutual Conductance (mA. per volt.)
R/F and Detector	20,000 to 50,000	15 to 50	0·7 to 1·5
A/F stage	4,000 to 12,000	10 to 30	0·9 to 1·8
Output valve	1,000 to 4,000	2 to 10	2·0 to 4·0

(ii) INDIRECTLY HEATED (A.C.) VALVES.

Purpose.	A.C. Resistance (Ohms.)	Amplification Factor.	Mutual Conductance (mA. per volt.)
R/F and Detector	8,000 to 20,000	15 to 85	3·0 to 6·5
Output valve	1,000 to 5,000	5 to 15	2·5 to 9·0

In earlier days in the Service, a multiplicity of valves was avoided in order to simplify storage conditions, and certain old sets are still in use in which so-called "general purpose valves" are to be found. It has not been possible to retain this simplicity, and more than 20 different receiving valves and nearly 50 different types of transmitting valves are now in use in the Service.

It should be noted that, with mass-produced valves, there are often considerable differences in the constants of valves of the same type. With the most careful precautions in manufacture the spread in characteristics from valve to valve may amount to $\pm$ 20 per cent. for certain constants, and $\pm$ 40 per cent. for others, such as g_m.

37. **Total Change in I_a due to Simultaneous Changes in V_g and V_a.**—The steady value of the anode current I_a is determined by the steady values of grid voltage V_g and anode voltage V_a. Suppose now that the grid voltage is changed by an amount v_g. If the anode voltage were kept constant, the corresponding change in anode current, i_a, would be given by

$$i_a = g_m v_g$$

provided the change were confined to the straight part of the appropriate mutual characteristic.

It has been seen, however, that unless special arrangements are made to keep it constant, the anode voltage will be affected by the change in grid voltage, owing to the different current flowing in the external circuit. This change in anode voltage in turn affects the nett change in anode current. Suppose that the anode voltage, either from the above reason or by a designed alteration in the H.T. supply, changes by an amount v_a. The change in I_a from this cause is connected with the change in V_a by the relation.

$$i_a = \frac{v_a}{r_a}.$$

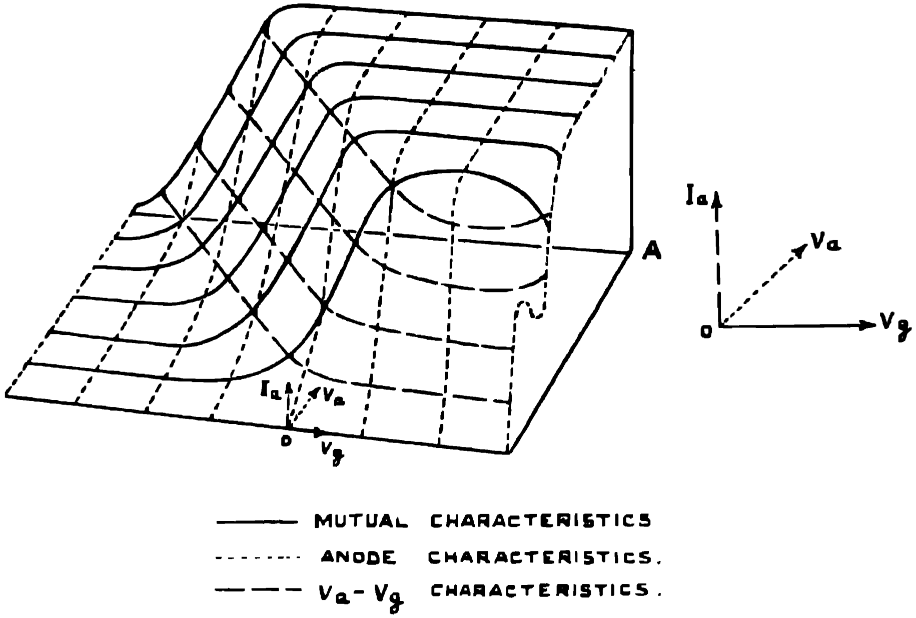
The total change in I_a due to simultaneous changes in V_g and V_a is therefore given by the sum of these two effects,

$$i_a = g_m v_g + \frac{v_a}{r_a}.$$

To prevent misunderstanding it may be added that v_a would be negative if due to an impedance in the external circuit. In the relation as given, the changes i_a, v_g and v_a are all supposed to be positive, or increases in the steady values

★38. The Characteristic Surface.—The relationship between the static values of grid voltage, anode current and anode voltage may be conveniently illustrated by the wire model shown in Fig. 22.

The model is in three dimensions, the (rectangular) axes being V_a, V_g and I_a. The origin of these is **not** at the point A, where three straight wires meet, but at the point O in one of the opposite sides. If we place the model as shown in Fig. 22 (*a*) these axes are viewed as shown in Fig. 22 (*b*), in which the positive directions only are indicated.



THE CHARACTERISTIC SURFACE OF A TRIODE.

(*a*) FIG. 22. (*b*)

(1) Looking along direction OV_a (Fig. 22) we see in outline the $V_g - I_a$ family of curves, *i.e.*, the mutual characteristics, for different constant values of V_a (Fig. 23). For negative values of grid voltage these rise in nearly parallel lines to a nearly common level X, the saturation value; but, when V_g becomes comparable with V_a, the grid takes more and more current and the curve begins to fall (at Y) without attaining saturation.

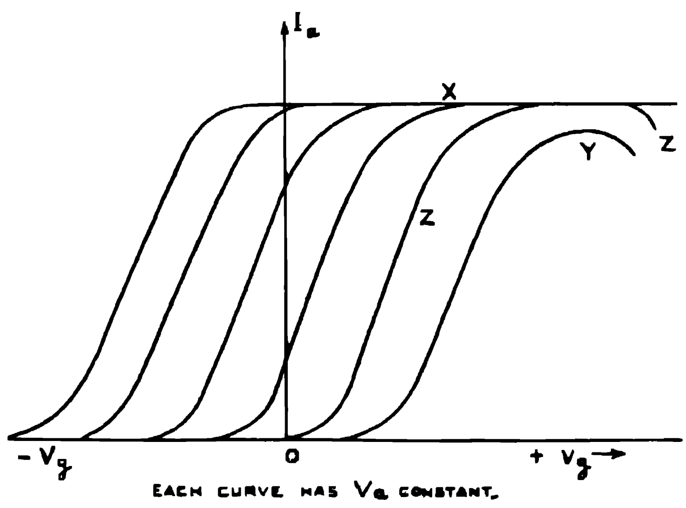


FIG. 23.

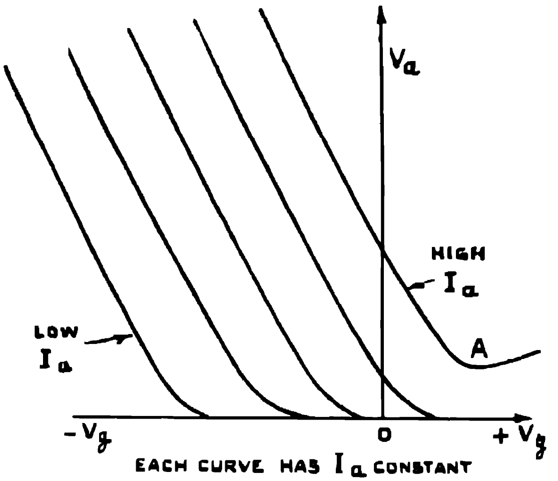


FIG. 24.

The other curves, such as Z, which have reached saturation will also commence to fall when V_g attains a value near the (constant) V_a of the curve in question, for the sum of the anode and grid currents remains approximately constant as long as V_a is constant, *i.e.*, along any one mutual characteristic.

(2) Looking vertically downwards on the model (Fig. 22) we see the $V_a - V_g$ system of curves for various constant values of I_a (Fig. 24). To maintain I_a constant, it is necessary to vary both V_a and V_g. As the negative grid bias is reduced it is necessary to reduce the value of V_a to retain constant I_a. When V_g becomes positive, however, and grid current commences to flow, V_a must increase again (as at A) to make up for the current taken by the grid. These curves are straight almost along their whole length, and so justify the generalised valve equation,

$$I_a = F\,(V_a + mV_g),$$

for along any one curve, I_a is constant, and so $V_a + mV_g$ is constant. This is the equation of a straight line connecting the variables V_a and V_g, the slope of the line being equal to m. These characteristics have some useful applications in the case of transmitting valves.

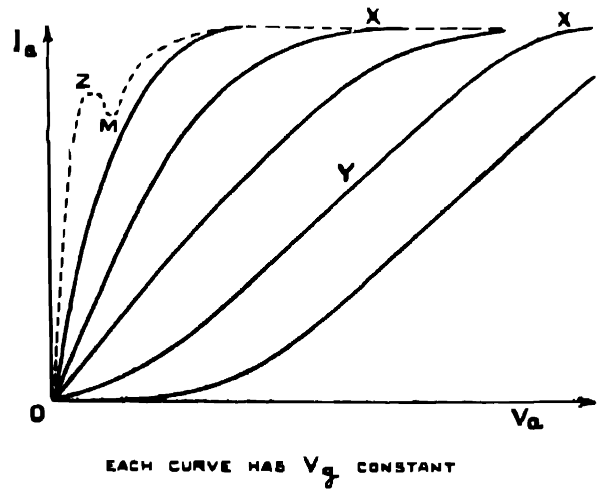


FIG. 25.

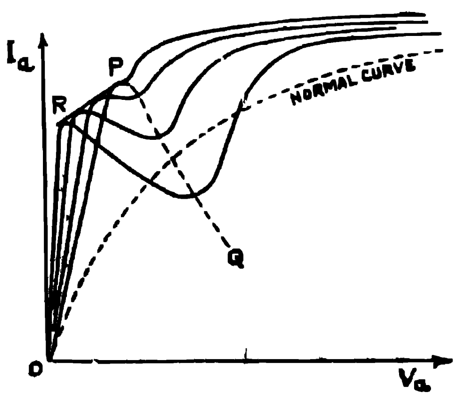


FIG. 26

ases in t.

(3) Looking *towards* the origin along V_gO we see the $I_a - V_a$ characteristics, *i.e.*, the anode characteristics for various constant values of V_g (Fig. 25).

Here again saturation is attained along XX for all values of V_a and V_g. (V_g is decreasing as we move to the right, so that OZX represents a positive, and OYX a negative, value of V_g). There is a curious reversal effect at Z, which is the commencement of a crevasse that becomes deeper and wider as we move upwards out of the sketch. It is explained as follows. With large positive values of grid voltage, and zero anode voltage, there is a grid current flowing. The bombardment of the grid releases many secondary electrons, which form a dense space charge round the grid. The smallest V_g will draw away a large number of these from the outer part of the cloud owing to repulsion from those inside ; and so I_a rises rapidly just at the start. The effect of raising V_a i to draw more primary electrons to the anode, and this reduces the grid bombardment. On this account, the copious supply of secondary electrons falls off and, as these are much in excess of the primaries, I_a will tend to fall. The two effects—increasing primary and decreasing secondary electrons to the anode—are in opposition and, owing to their greater number, at first the latter effect prevails, causing a dip in I_a. Ultimately, of course, as the anode voltage increases, the increasing number of primary electrons reaching the anode more than compensates for the decreased secondary emission from the grid, and I_a goes to saturation along XX, having previously fallen to a minimum at M With greater values of V_g, the effect is more marked and the crevasse grows larger. For clearness this is illustrated in a separate diagram (Fig. 26), with only one of the " normal " anode characteristics. The anode characteristics rise higher and higher as V_g increases, until the dip appears at P as described above. As V_g increases the dip becomes broader and deeper, forming a crevasse of which the base or col is PQ and the ridge PR.

39. **Dynamic Characteristics.**—It has been emphasised in the preceding paragraphs that the valve constants refer to static conditions and are indications of the suitability of valves for particular purposes, but that the results are modified when the valves are used under various circuit conditions. The ordinary characteristics described above are therefore usually referred to as "static characteristics," in contra-distinction to those which may be drawn to represent the behaviour of a valve under particular circuit conditions and which have been termed "Dynamic Characteristics." A useful example of dynamic characteristics has been given in paragraph 27, where the use of a "load line" (representing external circuit conditions) on the anode characteristics was described. Another form, and the one generally intended to be understood by the term "dynamic characteristic," is produced by drawing the load line on the mutual characteristics. Fig. 27 shows the circuit used, and a typical family of dynamic characteristics.

The discussion is similar to that employed in paragraph 27 for the anode characteristics. Suppose that the load in the anode circuit is R ohms. When the grid bias is sufficiently negative to prevent the flow of anode current, the potential of the anode is the same as that of the H.T. battery, which in Fig. 27 is 100 volts, and the conditions are represented on the family of mutual characteristics by the point A. Now if a current I_a is allowed to flow in the anode circuit, by reducing the grid bias, there will be a voltage drop in R of I_aR volts, so that the anode potential will be less than 100 volts by that amount. Suppose R is 40,000 ohms, and $I_a = 0{\cdot}25$ mA. The drop will be 10 volts, and the new conditions will be represented by the point C on the 90 volt characteristic where $I_a = 0{\cdot}25$ mA. and $V_g = -8{\cdot}5$ volts.

Now suppose that the grid bias is further reduced so that the anode current is still further increased to 0·5 mA. The voltage drop is now 20 volts (40,000 ohms × 0·5 mA.), and the new

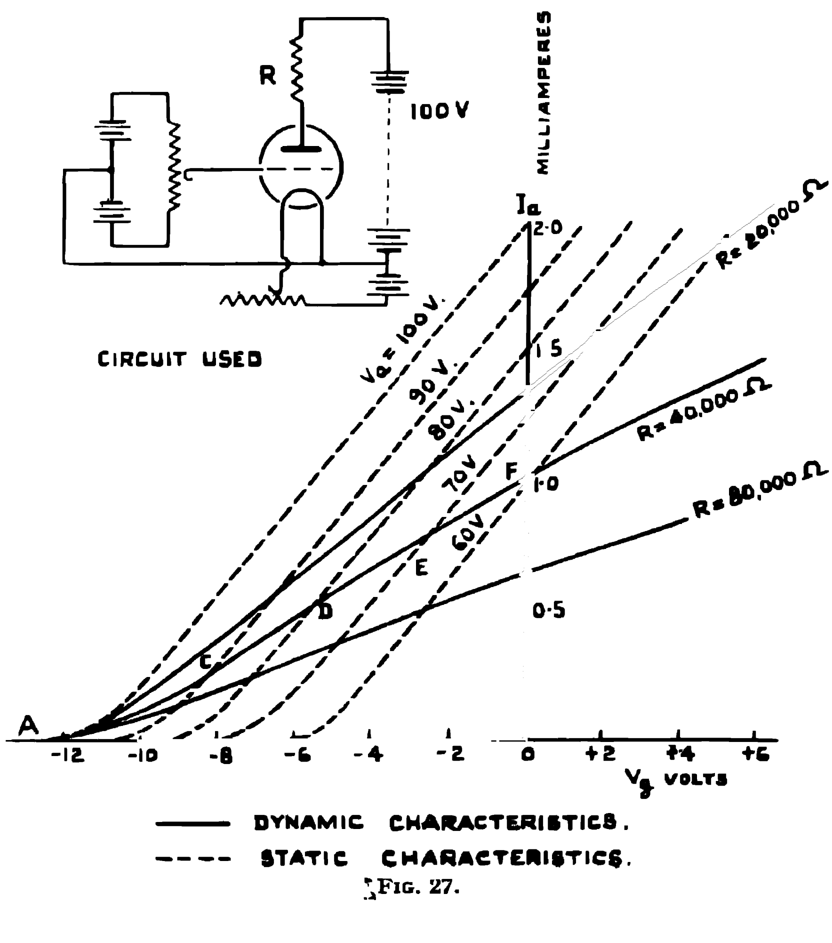


FIG. 27.

conditions are represented by a point D on the 80 volt characteristic, with grid bias approximately — 6 volts. The line joining the succession of points A C D E is the dynamic characteristic corresponding to the external load of 40,000 ohms. It is nearly straight except for a small portion at the extreme end near A, and from near this region to the point F where it cuts the zero grid voltage ordinate, it represents the range in which distortionless working is possible; equal alterations of grid voltage produce equal corresponding alterations of anode current. The figure also shows the dynamic characteristics for external resistances of 20,000 and 80,000 ohms.

It should be noted that the *dynamic characteristic* differs from the *load line* in that the former is only a straight line between certain points, provided that the mutual characteristics are parallel equally spaced straight lines over a portion of the graph.

If the external impedance contains a reactive component, the dynamic characteristic becomes a loop. It must be clearly understood that the dynamic characteristics are primarily characteristics of the circuit, and not of the valve alone. This subject will be further discussed when dealing with the valve as amplifier and transmitter (Sections "F" and "K").

40. **Inter-Electrode Capacity.**—The electrodes of the valves are conductors in more or less close proximity to one another, and each pair acts as a capacity. The actual values of the inter-electrode capacities are small, and are affected by the presence of electrons in the dielectric space; the influence of the latter depends upon whether the valve is being operated in a static state with the filament lit, or as a component of an oscillating circuit.

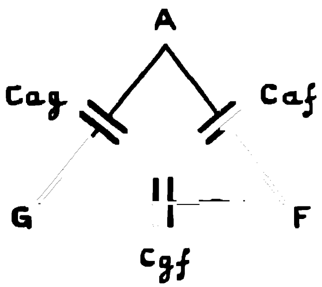


FIG. 28.

The effects of inter-electrode capacity are of considerable importance, and will be described in the appropriate sections of this Handbook; in this place, only the nature, extent, and variations of the property are being considered.

Not only are the electrodes themselves responsible for the apparent inter-electrode capacities, but their supporting wires, and the leads to the external contacts, enter very largely into the problem. The method of assembling the electrode system on wires secured in a pinch, invariably used for receiving valves and described earlier in this chapter, is responsible for a considerable addition to the true inter-electrode capacity, and it is therefore avoided as far as possible in valves made expressly for operating on very high frequencies, such as "**acorn type**" valves.

The capacities between the three electrodes of a triode may be represented by the diagram of Fig. 28. A, G and F represent the anode, grid and filament respectively, and C_{ga}, C_{gf}, C_{af} represent the capacities between the three electrodes. It is obvious that when considering any one of these, the effect of the other two in series cannot be neglected. The capacities are usually measured by a bridge method using A.C. of (say) 500 or 1,000 cycles/second, with telephones as a balance indicator Three observations are taken with the electrode terminals connected in pairs successively to eliminate one of the capacities. Thus with G and F connected together, the first observation (i) $= C_{ga} + C_{af}$. With A and F connected together the next observation gives the "**input capacitance**" (ii) $= C_{ga} + C_{gf}$; and with A and G connected together, the last observation (iii) $= C_{af} + C_{gf}$.

Hence (i) + (ii) — (iii) $= 2C_{ga}$.
(i) + (iii) — (ii) $= 2C_{af}$.
(ii) + (iii) — (i) $= 2C_{gf}$.

The values obtained for the capacities of a small receiving valve, when measured in this manner, obviously include the capacities between the metal contacts of the socket in which the valve is placed for the test, and due allowance must be made for these. Similarly, the capacities of the flexible leads of transmitting valves must also be taken into consideration.

The approximate values of the inter-electrode capacities of certain types of valve are given in the following table :—

—	C_{ga}.	C_{gf}.	C_{af}.
	μμF.	μμF.	μμF.
Small receiving valves	2·0 to 8·0	2·5 to 4·5	2·5 to 3·5
Glass transmitting valves—			
150 watts	5	8	3
450 watts	13	5	9
Silica valves—			
2·5 kW.	11	9	7
4·0 kW.	14	11	9
15·0 kW.	26	20	10
Acorn valves	1·2	0·9	0·45

Variations in inter-electrode capacity are produced by changes in the space charge between the electrodes, or alterations in density of the moving electrons.

In the **grid-anode space,** the electrons are moving more and more rapidly as they approach the anode, and, therefore, the electron density is decreasing towards that electrode. The total current flowing across this space is the sum of the electron current plus the "capacity current" between anode and grid. Moreover, this total current must have the same value at all planes parallel to the electrodes, since we know that current cannot accumulate at any point in space (Kirchhoff's Law). Now the electronic current at any point is proportional to the electronic density and velocity at that point ; since the density *increases* as the grid is approached, it is clear that the capacity current must *decrease* in order to maintain the total current constant. This reasoning implies that the stored electric energy must decrease towards the grid and, if it were summed up over the inter-electrode space, it would be less than if no moving electrons were present. Equating this reduced total energy to $\frac{1}{2}C_{ga}V^2$, it is evident that a decrease in energy for the same applied voltage must mean a decrease in effective capacity C_{ga}. The **grid-anode capacity** will therefore **decrease** during each alternate half cycle when a pulse of anode current flows past the grid.

An alternative argument leading to the above conclusion is provided by the inertia of the electrons. The movement of an electron lags slightly behind the cause which produces it. This "lagging current" may be attributed to an inductance acting in series with a resistance, the whole being in parallel with the anode-grid inter-electrode capacity. This equivalent inductance is very small, but its inductive reactance operates in anti-phase to the capacitive reactance of the inter-electrode capacity, and slightly reduces the value of the latter when a pulse of anode current flows past the grid.

In the **grid-filament space** there is a second effect which comes into operation and overcomes the phenomenon described above, and gives an increase in this capacity when electron current flows, provided that the current is "space charge limited." This new effect is due to the fact that the hot filament shoots off electrons with appreciable velocity, and over a small space around the filament (or cathode) the electron density is so great that the voltage gradient is reversed, and most of the electrons return to the filament. This means, virtually, that the filament can be considered as having a diameter equal to the distance from the filament to the surface of zero voltage, or the place where the mean velocity of the electrons is a minimum. The **grid-filament capacity** is measured from the grid to this surface (which may be a plane), and **is greater** than if no electrons were present, since the "plates" of the condenser are closer together. This effective increase in diameter of the filament also increases the mutual conductance of the valve.

Recent work has shown that the grid-filament capacity increases with anode current up to the point at which grid current flows. Moreover, the increase in C_{gf} is greater the higher the filament temperature, but is not a simple function of any of the variables concerned. The decrease in C_{ga} is much smaller proportionately than the increase in C_{gf}, the proportion being about 1 : 10.

When the grid of a valve is biased negatively, or during the negative half cycles if the valve is oscillating, the negative field at the filament, due to the high negative potential on the grid, prevents the electrons from being shot out to so great a distance as when the grid is more positive. The effective diameter of the filament is thus less than it would be if the grid were more positive, or during the positive half cycles in the case of an oscillating valve.

It is clear that the grid-anode capacity will cause changes of potential on the anode to react on the grid. In other words, the grid-anode capacity acts as a direct capacitive coupling device between the output and input circuits. As the potential changes on the anode are primarily caused by potential changes on the grid, it follows that back coupling can cause instability if the circuit conditions allow it (Section "F"). For example, if the potential changes fed back to the grid are in phase with the input variations, there will be **regeneration,** and the valve which is only required to *amplify* oscillations, will *produce* oscillations instead. This is most marked in valves of high amplification. Back coupling also throws a load on the input circuit, and is responsible for small power losses, essentially of the same nature as those produced by grid current (Section "F").

The instability of an R/F amplifying valve has long been a well-known phenomenon, and several methods have been used to counter it. In all cases, these have aimed at either neutralising the *effect* of the grid-anode capacity, or reducing the capacity to the smallest possible amount. The various methods of neutralising are discussed in Sections "F" and "K"; subsequent paragraphs in this section describe the means which have been adopted to minimise this capacity by special valve construction (screen grid valves and R/F pentodes).

(a)

(b)

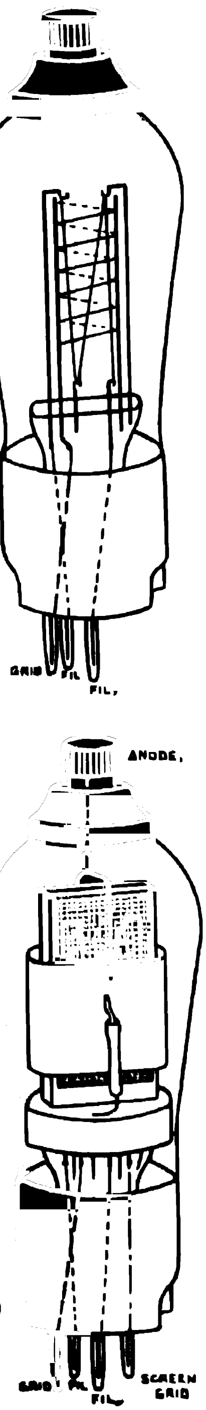


FIG. 29.

41. **The Screen-Grid Valve.**—This valve was originally introduced to overcome ill effects due to anode-grid capacity in R/F amplifying stages. It has two grids between filament and the anode; the grid near the filament performs the same function as the grid of a triode and is therefore frequently referred to as the "**control grid,**" while the second grid acts as an electrostatic screen between the control grid and the anode and is therefore termed the "**screen-grid,**" or simply the "screen."

The construction of a screen-grid valve is shown in Fig. 29 (*a*) and (*b*). Inside the bulb is a flattened cylindrical anode connected to a terminal cap on the top of the bulb. It is supported by a short glass rod on the flange or skirt of metal which is attached to, and electrically forms part of, the screen-grid, which can be seen as a close mesh inside the anode. The control grid and cathode are completely enclosed within this screen-grid assembly.

The magnesium getter, invariably used in the manufacture of small valves, is placed on the anode and consequently deposits only on the glass above the flange of the screen. Sometimes a light wire welded on the screen presses on the interior of the bulb so as to put the magnesium film in contact with the screen, and to make the screening of the whole more effective. For clearness, the wire and magnesium getter are not shown in the figure. The reason for the interposition of the screen will be fully understood after the section dealing with

the valve as an amplifier has been read, but it may be stated here that it is a device to reduce the grid-anode capacity to the lowest possible value.

The screen is kept at a high potential approaching that of the anode. Its functions may be best understood by considering the lines of force between the electrodes. If the screen were a sheet of metal instead of a mesh or grid, it is clear that all lines of force from the control grid and filament would end on the screen, and there would be no capacity between the control grid and the anode. It may be considered that with the screen an open mesh and nearly at the same potential as the anode, practically all the lines of force from the grid and filament end on the screen. When the potential of the screen is below that of the anode there is a field between them, and a few lines also join the grid and anode, *i.e.*, there is a slight residual grid-anode capacity. Although the majority of the lines of force from the grid end on the screen, the large majority of the electrons travelling towards the screen, from the vicinity of the grid, are projected by their momentum through the spaces of the mesh, and so come under the screen-anode field and are collected by the anode. Thus the screen grid valve can function in the same manner as a triode, though its characteristics are peculiar to itself. In commercial types of screen grid valves the residual grid-anode capacity is from 0·001 μμF. to 0·02 μμF., while the corresponding value for triodes is about 2 to 8 μμF.; the screening effect of the fourth electrode is therefore very pronounced.

In order to have the screening as complete as possible it is also necessary to separate the grid and anode circuits external to the valve by a similar electrostatic screening device. For this purpose it was formerly the practice to insert the valve in a hole in an earthed metal screen, supported in the same plane as the internal screen electrode (*cf.* F.34, Fig. 25). With the introduction of metallised valves, the separation of anode and grid circuits has become somewhat easier, and it is now usually the custom only to have a small metal cup to shield the anode, if it is at the top of the valve.

It will be clear from the above discussion that slight changes in anode voltage have practically no effect on the *field* between the grid and the screen, which is responsible for driving the anode current across this region. In other words, the impedance or A.C. resistance of the valve is very high. The grid control, however, is obviously practically the same as if the screen and anode together formed a collecting electrode as in a triode; *i.e.*, the mutual conductance is of the same order as for a triode. Hence, from the relationship between the valve constants (paragraph 35), the amplification factor is high. The following table of practical values may be compared with that given for triodes in paragraph 36; as in the case of triodes, the valves with indirectly heated cathodes have the highest amplification factors.

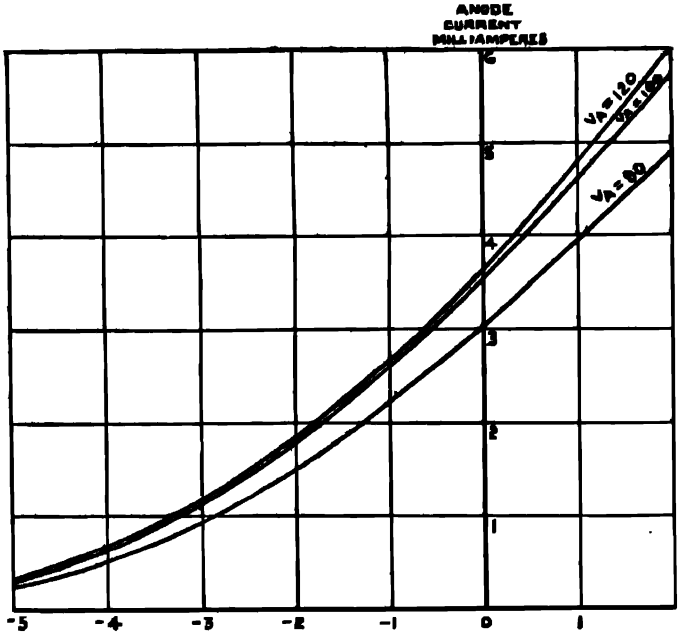
CONSTANTS OF SCREEN GRID VALVES.

A.C. Resistance (Ohms.)	Amplification Factor.	Mutual Conductance (*m*A. per volt.)	Grid-anode Capacity (μμF.)
200,000 to 1,000,000	100 to 1,500	0·5 to 3·5	0·001 to 0·02

(H.T. voltage, 120 to 150 volts; screen voltage, 70 to 90 volts.)

In order to take the fullest advantage of the high amplification factor of the screen grid valve, it is essential that the anode circuit should also have a very high impedance. This will be clear from a consideration of the discussion on the load line in paragraph 27. Unless the anode circuit impedance is very high, the changes in anode current brought about by changes of grid voltage will not produce the desired changes in anode potential.

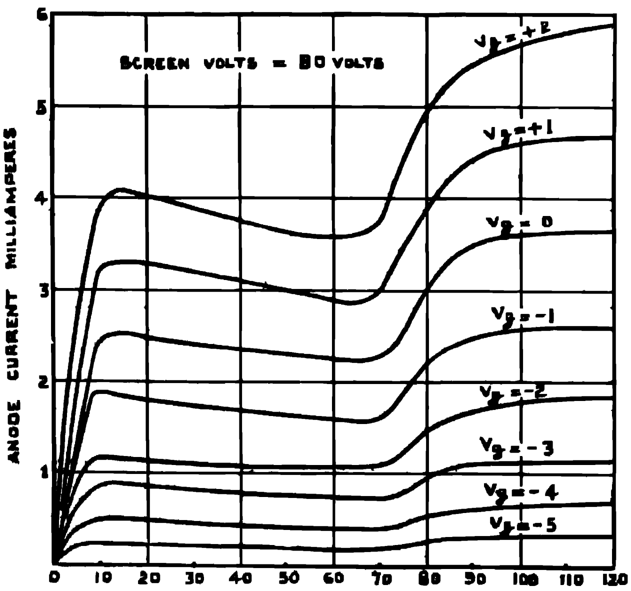
The mutual characteristics of the screen grid valve (Fig. 30) are similar in shape to those of the triode, but the curves for different values of anode voltage are closer together, indicating the high A.C. resistance. Obviously, the only valuable curves are those for anode voltages higher than the screen voltage.



MUTUAL CHARACTERISTICS OF A SCREEN GRID VALVE.

SCREEN VOLTS = 80.

FIG. 30.

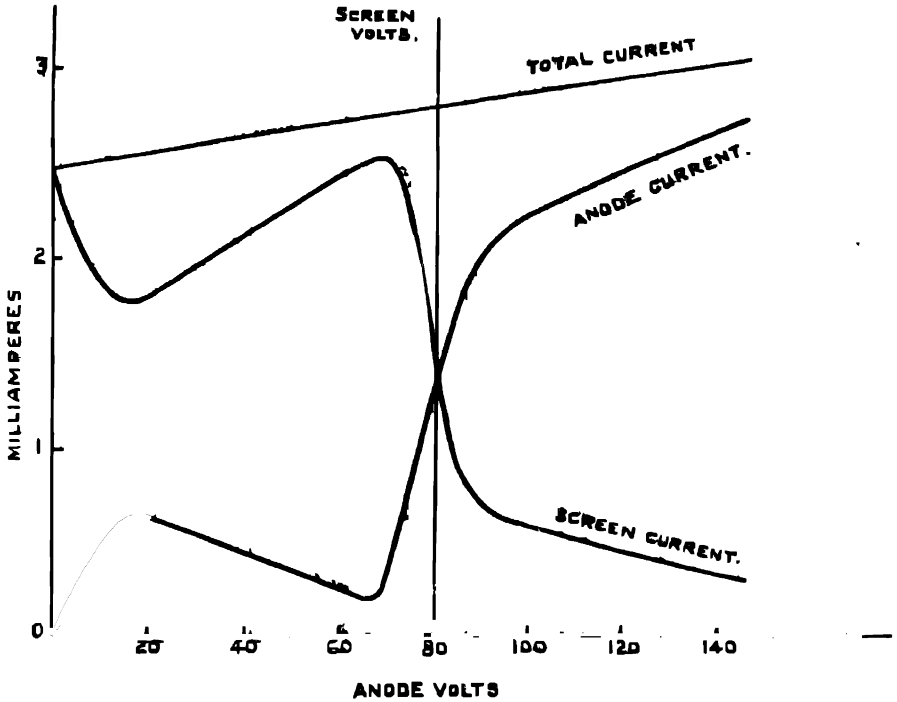


ANODE CHARACTERISTICS OF A SCREEN GRID VALVE.

FIG. 31

The anode characteristics (Fig. 31) are very different from those of the triode; as the anode voltage is raised from zero, keeping the screen at the rated value for the valve, the anode current rises until in the vicinity of 12-20 volts it attains a maximum value and then falls. This fall is explained by the emission of secondary electrons from the anode, due to the impact of the primary electrons. The secondary electrons travel to the screen, which is at a higher potential than the anode. The number of secondary electrons emitted by the anode depends on the number of primary electrons and on the voltage through which they have fallen on reaching the anode (*i.e.*, on their velocity). If the screen voltage is made high enough, it is actually possible to produce a current of secondary electrons from the anode larger than the primary current, *i.e.*, the curve falls below the anode voltage base. With ordinary adjustments, the anode current falls as the anode voltage is raised until the latter approaches the screen voltage. Then the reverse condition occurs. The primary electrons impinging on the screen produce secondary electrons which travel to the anode, while those emitted from the anode return to it and are not detected.

Fig. 32 shows the variations of screen and anode current with anode voltage, the screen voltage being fixed at 80 volts, and it also shows the rising total current, which is the actual value of the current passing through the control grid plus secondary emission currents. When the valve is in use as a radio frequency amplifier, it is obvious that to avoid effects due to secondary emission from the anode, the adjustment of potentials must be made so as to produce conditions corresponding to the nearly horizontal portions of the anode characteristics, *i.e.*, when the anode voltage is higher than the screen voltage. The best adjustment is when the screen voltage is about two-thirds of the anode voltage. It may be obtained either by a tapping on the H.T. battery, or by inserting an appropriate resistance of some 50,000 ohms



SCREEN VOLTS = 80
GRID VOLTS = 0
ANODE AND SCREEN CURRENTS PLOTTED AGAINST ANODE VOLTS.

FIG. 32.

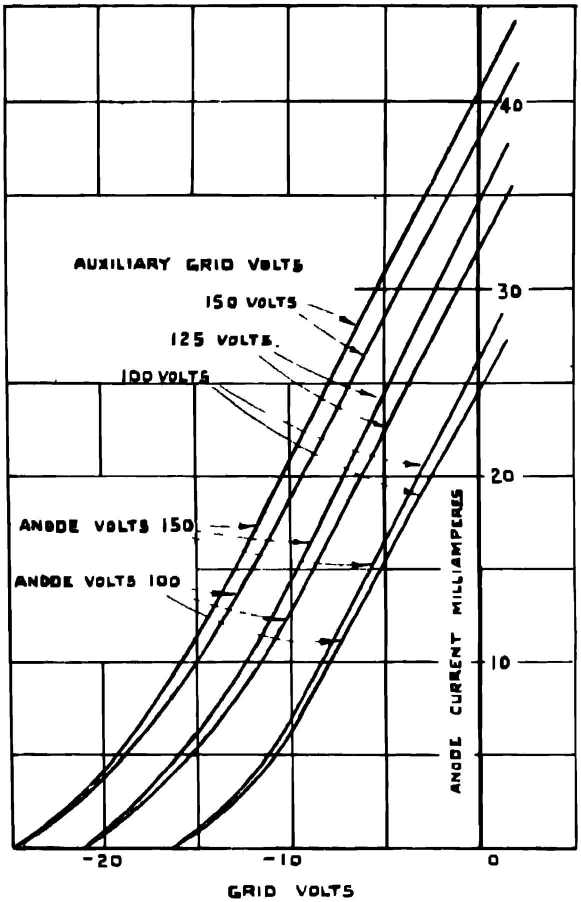
or so between the screen and H T positive. The value of the resistance depends on the screen current, which is about 0·5 *m*A in the majority of screen grid valves operating with 120-150 volts on the anode.

Screen grid valves came into general use after Hull's work in 1926, but the initial idea was due to Siemens and Halske in 1916.

42. **The Pentode. Output and R/F Pentodes.**—A five-electrode valve, the pentode, is to be regarded as an improvement on the screen grid valve for specific purposes. It was originally introduced for the last or power output stage of receiving circuits, but has since become widely used in R/F amplifying stages, and to a smaller extent for detection. Moreover, **transmitting pentodes** are becoming increasingly popular ; the small value of the inter-electrode capacity often makes it possible to eliminate neutralising condensers and to simplify the design.

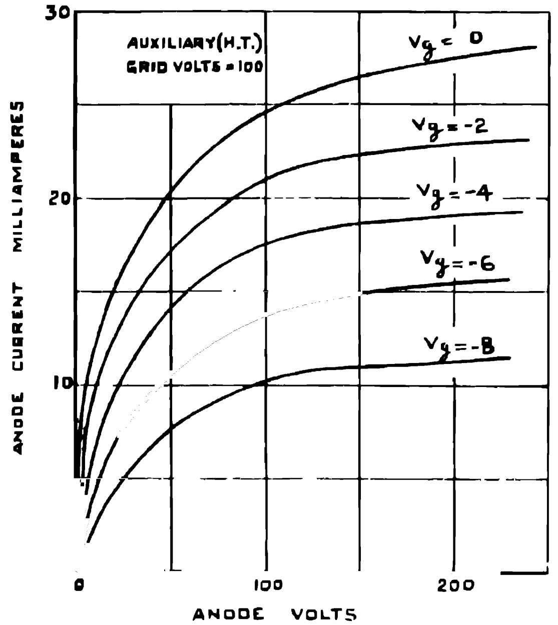
The valve has three grids between the cathode (or filament) and the anode. Numbering from the filament, the first grid is the usual **control grid,** the second is termed the auxiliary or **priming grid** and is externally connected to a source of high potential, the third is called the **suppressor grid,** and in A/F power valves is earthed by internal connection to the filament or cathode, but is taken to a separate terminal or base pin in the case of R/F pentodes. The lead from the auxiliary grid is connected either to a screw terminal on the side of the base, or to a pin situated in the base.

The function of the earthed grid in output valves is to suppress the emission of secondary electrons, either by the anode or the auxiliary grid. This function is clearly understood when it is realised that secondary emission is only important when it can be collected by an adjacent electrode at a higher potential than the emitter.



MUTUAL CHARACTERISTICS OF PENTODE.

FIG. 33



ANODE CHARACTERISTICS OF PENTODE

FIG. 34.

By the suppression of secondary emission one is able to eliminate the disadvantages associated with the "negative resistance kink" always found in the anode characteristics of a screen grid valve. It may be noted here that the **dynatron oscillator** makes use of the negative resistance feature (the anode current *decreases* as the anode volts *increase*) referred to above. Moreover, the form of the pentode characteristics makes it suitable for use as an output valve.

When the pentode is designed for **R/F amplification,** the second grid (H.T. or priming grid) is constructed as a complete electrostatic shield, as in the case of the screen grid valve. When, however, the valve is designed as an **A/F power output** valve, this elaborate form of electrode construction is not necessary, and a grid of simpler and cheaper construction is used.

Figs. 33 and 34 show the characteristics of a typical commercial pentode ; the anode curves of Fig. 34 are seen to be similar to those of the tetrode (Fig. 31), but without the secondary emission kink. The constants of commercial types of pentodes are of the following order.

CONSTANTS OF PENTODES.

A.C. Resistance (Ohms.)	Amplification Factor.	Mutual Conductance. (mA per volt.)
30,000 to 100,000	50 to 100	1·3 to 4·0

43. **Frequency Changer Valves. Hexodes, Heptodes and Octodes, etc.**—When the pentode was first designed, the suppressor grid was operated at earth potential, permanent connection to the filament usually being made within the valve. The advent of **suppressor grid modulation** (N.24) showed that variations in potential of the suppressor grid produce corresponding alterations of output power following almost a straight line law. As the potential of the suppressor grid is made more negative, the anode current is reduced, and, at the same time, the anode impedance is reduced. If the suppressor grid has a very open mesh, this control of the anode current (**modulation**) is relatively weak ; if the suppressor grid were of closer mesh the control would be stronger, a basic idea accounting for the development of hexodes, heptodes, and octodes, which were **designed for the control of the anode current by two separate grids.**

The rapid growth in popularity of the superheterodyne type of receiver since 1917, led to an insistent demand for a simple " **frequency changer,**" to change all incoming R/F signals to a common intermediate frequency for the purposes of amplification. Since 1929, single valve frequency changers have developed considerably, and now appear gradually to be replacing the two valves always previously required for this purpose. Their circuits and principles of action differ considerably, but all aim at " **frequency multiplication** " (N.23) ; this involves the generation of several frequencies in the anode circuit of a valve, among which are the sum and difference of the oscillator and signal frequencies respectively The **difference frequency** is selected by the tuned secondary circuit of a transformer coupling, and constitutes the **intermediate frequency** which is passed on for subsequent amplification.

HEXODES.—One form of hexode is a modified pentode, having an extra electrostatic screen between the suppressor grid and the anode. The signal voltage is applied to the inner grid and the oscillator voltage, obtained from a separate oscillator, is applied to the outer (suppressor) grid. The signal input is " mixed " with the oscillator frequency by electron coupling, and the difference frequency (intermediate frequency) is selected by the anode circuit.

A more usual form of hexode is a valve in which the first grid is the signal input grid. This is followed by an electrostatic screen at positive potential, the oscillator grid, the oscillator anode, and the output anode. Again, the signal input is mixed with the oscillator frequency. The simple hexode never enjoyed great popularity, since it was never possible to apply A.V.C. to the signal input grid ; the oscillator derives its electron supply from a " **virtual cathode** " situated near the electrostatic screen, and if A.V.C. were applied to the signal grid the oscillator usually stopped oscillating.

HEPTODES (Pentagrids).—The first form of heptode had the features of the modified pentode described above, but had a suppressor grid between the extra screen and the anode. This form of heptode is used with a separate oscillator valve to provide the modulating signal, and the arrangement of the electrodes is shown below in the schedule for Type (*a*) heptodes. The valve may be assumed to have concentric cylindrical construction, and the electrodes are numbered from within, starting with the cathode. It may be noted that this type of heptode still exemplifies the basic principle of suppressor grid modulation.

There is a newer and more commonly used type of heptode which incorporates the function of oscillator valve and modulator valve within a single valve envelope, an **electron coupling** being used to mix the two frequencies. In this case, the arrangement of the electrodes consists of five grids (**pentagrid**) inserted between the cathode and anode ; the order of these grids is **shown** below in the schedule for Type (*b*) heptodes. In this case, also, concentric cylindrical valve construction may be assumed, the first two grids and cathode constituting the oscillator portion of the valve ; the remainder of the valve constitutes a tetrode, the whole being virtually an **electron coupled triode-tetrode combination.** The tetrode is sometimes called the modulating portion of the valve, and the signal grid (5) is usually given variable mu characteristics.

When the oscillator grid is negative the anode current is reduced, and when the oscillator grid is positive, the anode current increases. Every electron which reaches the modulator has to pass through the oscillator portion of the valve. Clearly, if the grid of the latter is oscillating/

the anode current will rise and fall at the frequency of the oscillator, frequency multiplication thus being achieved by electron coupling. This matter is referred to again in F.51, a heptode circuit being given in N.63—Fig. 41.

ARRANGEMENT OF ELECTRODES IN HEPTODE VALVES.

	Type (*a*)		*Type* (*b*).
7	—Anode.	7	—Anode.
6	—Suppressor grid.	6	—Second screen.
5	—Second screen.	5	—Signal grid (variable mu).
4	—Oscillator grid (Separate oscillator).	4	—First screen.
3	—First screen.	3	—Oscillator anode.
2	—Signal grid.	2	—Oscillator grid.
1	—Cathode.	1	—Cathode.

OCTODES.—Two types of octode have been developed to supersede the heptode. The **suppressor grid octode** is a Type (*b*) heptode, having a suppressor grid between the second screen and the anode. The whole, therefore, consists of an electron coupled triode-pentode combination.

The **velogrid octode** dispenses with the earthed suppressor grid, having instead a positive sixth grid termed the " velogrid " which provides accelerator action.

At first it appeared that the octode might become the most used " mixer valve," but in 1935-36 it was observed not to be entirely satisfactory, owing to a falling off in the conversion conductance (paragraph 44) towards the higher frequencies. Investigation arising out of this trouble led to the " triode-hexode."

TRIODE-PENTODES.—This might be regarded as a multiple valve, since it consists of a triode oscillator and a pentode included, for convenience, within one valve envelope; it is also, however, a frequency changer. Electron coupling is not employed, and mixing is achieved by inserting the oscillator frequency into the signal frequency circuit by coupling external to the valve. The valve was developed in 1934, and represents the best features of simple additive mixer systems.

TRIODE-HEXODES.—This valve consists of a triode oscillator and a hexode, both contained within the same valve envelope. The hexode portion of the valve is spaced to one side of the triode and, starting from the cathode, has electrodes in the order—input grid, screen grid, injector grid, second screen grid, and the output anode; the injector grid is linked to the grid of the triode. It represents the peak of development in single valve frequency changers employing electron coupling.

44. **Conversion Conductance (or Conversion Transconductance) g_t.**—This is a new term which has arisen in connection with the use of frequency changer valves, the object of which is to receive a signal input at *one frequency* and to give a total anode current variation which contains a component varying at *another frequency* (**the intermediate frequency**). This constant may be measured in various ways not detailed here, and could be represented by the symbol g_t; it is similar to mutual conductance g_m, and may also be expressed in milliamps of current at intermediate frequency per volt of signal on the grid. It is a quantity which depends upon the efficiency of conversion, as well as on the amplifying properties of the valve. A knowledge of the conversion conductance makes it possible to compute the amplification produced by a heptode at the intermediate frequency, in just the same way that g_m enables one to estimate the V.A.F. of an SG tetrode amplifying stage. In Section " F " it is shown that the V.A.F. of the latter is approximately given by g_mZ, where Z is the impedance of the external circuit at the frequency in question; in a similar manner, the effective amplification of a frequency changing stage may be written as g_tZ.

In practice it is found that the value of g_t is approximately half that of g_m, many modern frequency changers having a value of 0·5 mA/volt for g_t. The above relation is only true if the

conversion conductance is measured with the oscillator grid at the most positive voltage to which the oscillation carries it. Moreover, it is essential that the oscillator input should be adjusted to its optimum value ; too high an input, or too low an input, reduces the gain of the valve.

Sometimes values of g_t are expressed in " reciprocal ohms," the above numerical value then being 500 micromhos.

45. **Multiple Valves. Miscellaneous Valves.**—With reference to paragraph 2, it will be observed that all multi-electrode valves may be considered to be derived from simple types, and in this work they are classified into frequency changer valves, multiple valves, and valves used for miscellaneous purposes.

Multiple valves are those which do not introduce any essentially new principles, and are merely combinations of simple valves assembled within one envelope. These include double-diodes, double-diode-triodes, double-diode-pentodes, double output valves (triodes and pentodes), etc. The requirements of automatic volume control have been responsible for the development of some of these multi-purpose valves ; they have the advantage of saving a considerable amount of space.

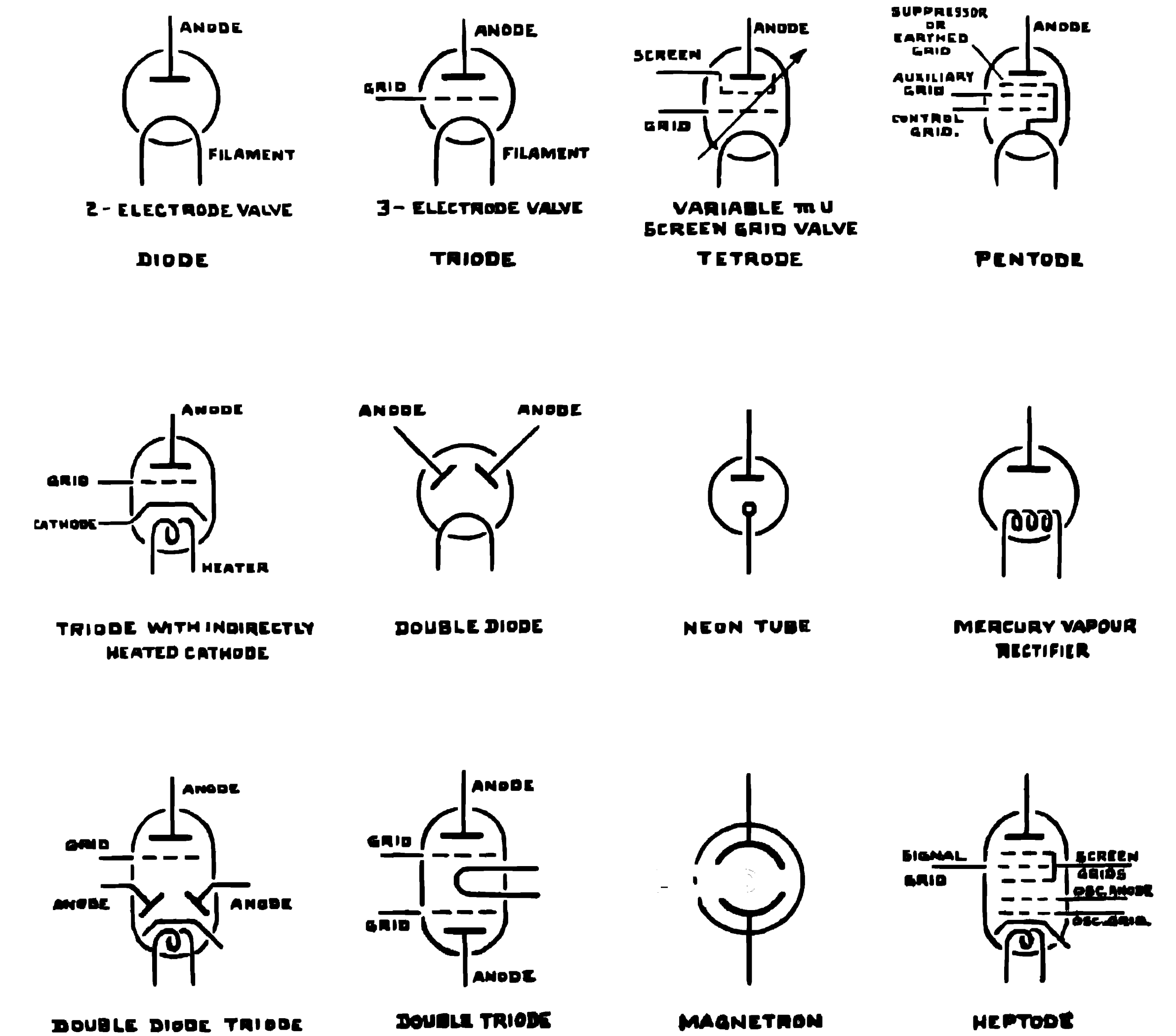


FIG. 35.

Among valves used for miscellaneous purposes, one must include the magnetron, which is *essentially* a cylindrical diode with a split anode. It is described in detail in K. 60-65.

Other miscellaneous valves include the thyratron, a soft triode having a little mercury vapour within it.

More recently another type of valve has been developed by Zworykin, which depends for its action on the production of secondary electrons (secondary emission); these valves are usually known as " electron multipliers," and, after further development, they may find some application as amplifiers.

No treatment of thyratrons or electron multipliers is included in this work.

46. **Valve Symbols.**—The symbols used throughout this book to represent the various types of thermionic valve, are shown in Fig. 35.

THERMIONIC VALVES.

EXAMINATION QUESTIONS.

NUMERICAL EXAMPLES.

1. The following readings were taken on a three electrode valve. Plot the characteristic grid voltage-anode current curves, and determine the A.C. resistance at zero grid volts, the amplification factor, and mutual conductance.

Grid Voltage.	0	−2	−4	−6	−8	−12	−14	−16	−18
I_a in mA— (a) with 130 volts H.T. .. .	15	13	11	9	7	3	1·5	0·7	0·2
(b) with 100 volts H.T.	10	8	6	4	2	0·4	0·1	—	—

(Results: A.C. Res. = 6000 ohms; m = 6; g_m = 1mA/Volt. C. & G. Grade 1., 1932).

2. In a certain triode it was found that a current of 5 mA was obtained when the anode and grid potentials had the following values:—

V_a	500	310	139	86	58	35	10
V_g	−32	−125	3·5	9·5	13·5	16·5	26

Plot the curve relating V_a and V_g for this anode current, and deduce the amplification factor of the valve. Discuss why this factor becomes smaller when the grid is very positive.

(Result: 10.7. I.E.E. May, 1932.)

3 A three electrode thermionic valve has a total emission of 15 mA, an amplification factor of 10, and an A.C. resistance of 30,000 ohms. It is further known that when the grid is joined to the negative end of the filament, the anode current with an anode voltage of 100, is 7·5 mA.

Draw a family of characteristic curves at anode voltages of 200, 150 and 100.

(C. & G. Final, 1925.)

SECTION "B."

4. The relation between anode current and anode potential of a triode valve, in which the grid potential was zero, is given in the following table :—

Anode potential in V ..	50	100	150	200	250	300	350
Anode current in mA	1·0	3·5	6·5	9·3	10·1	10·2	10·2

The amplification factor of the valve was 7 0.

Derive the characteristic relating anode current and grid potential for a constant anode potential of 150 V.

(I.E.E. November, 1932.)

5. The following values of anode current were obtained with a triode :—

Anode Voltage.	25	50	75	100 volts
I_a with $V_g=0$	0·4	2·8	6	8·5 milliamps
I_a with $V_g=-4{\cdot}2$	0	0·6	3·0	5·7 milliamps
I_a with $V_g=-8{\cdot}3$	0	0	1·0	2·9 milliamps

Plot the I_a—V_a characteristic curves.

If a battery of 150 volts, and a resistance of 15,000 ohms are connected in series with the anode and cathode, what will be the I_a at the above three values of grid voltage?

(Result : 5·4 mA ; 4·3 mA ; 3·1 mA. C. & G. Prelim., 1935.)

6. A high-mu triode valve has the following characteristics :—

Plate Voltage.	Plate current in milliamperes at following grid potentials.						
	−3·0	−2·5	−2·0	−1·5	−1·0	−0·5	0
60	0	0	0	0	0·08	0·4	0·9
80	0	0	0	0·01	0·19	0·59	1·16
100	0	0	0	0·06	0·30	0·79	1·45
120	0	0	0	0·12	0·49	1·02	1·76
140	0	0	0·03	0·24	0·68	1·28	—
160	0	0	0·10	0·40	0·92	1·52	—
180	0	0·03	0·18	0·58	1·16	—	—
200	0	0·08	0·30	0·80	1·41	—	—
220	0·02	0·14	0·48	1·05	—	—	—
240	0·05	0·24	0·68	1·30	—	—	—

Plot the characteristics with the plate voltages as abscissæ and the plate currents as ordinates for the given values of grid potentials.

If an external resistance is inserted in the plate circuit, what will be the plate currents with an H.T. battery of 250 volts and grid voltages of —2·0, —1·5, —1·0, —0·5 and 0, when the external resistance is (*a*) 100,000 ohms, (*b*) 500,000 ohms ?

(Result : 0·4, 0·15 ; 0·64, 0·22 ; 0·9, 0·3 ; 1·17, 0·38 ; 1·47, 0·46. C. & G. I., 1937.)

SECTION "B."

7. *Explain why the anode current of a thermionic valve may be limited by the filament current, and also by the presence of a space charge. How would you measure the amplification factor of both a small receiving valve and of a high-powered high-voltage transmitting valve?*

(I.E.E. May, 1931.)

8. Account for the very high amplification factor which is obtained by the use of a screen grid valve. Why is secondary emission of importance in the action of this valve, and how does this limit the magnitude of the grid swings with which the valve can deal? Briefly explain how a pentode will overcome this restriction.

(Qual. Wt. Tels., 1936.)

9. Sketch, diagrammatically, a heptode frequency changer. Explain its action fully. Sketch and describe its associated circuits.

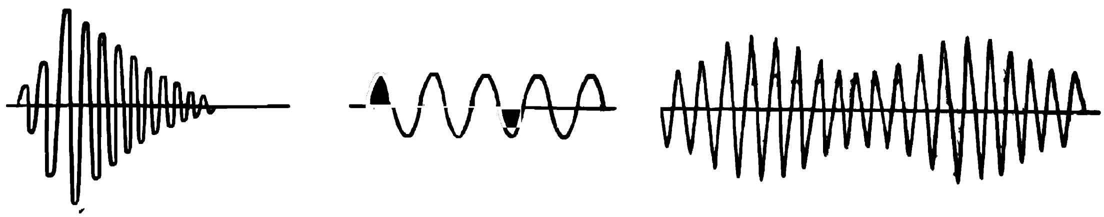
(Qual. Wt. Tels., 1935.)

10. Sketch a typical series of I_a/V_a curves for an output pentode valve. Draw in a load line to represent "optimum loading," explaining the meaning of the term.

(Qual. Wt. Tels., 1934.)

RECEPTION OF ELECTROMAGNETIC WAVES.

1. **The Raw Material of W/T Communication.**—We have seen in Section " A " how radio-frequency currents, having the form of damped wave trains [Fig. 1 (*a*)], are generated in the aerial circuit of the transmitting station by means of spark circuits. Use of damped wave trains is,



(a) (b) (c)

FIG. 1.

however, obsolescent, and Fig. 1 (*b*) and (*c*) represents the wave forms known as **continuous wave** (C.W.) and **interrupted continuous wave** (I.C W) respectively, both of which are in common use. A more exact and international classification of wave forms is given in **K.51**, but for simplicity, at this stage, it is sufficient to appreciate that W/T communication may be achieved by means of R/F currents having the wave forms shown in Fig. 1.

These radio-frequency currents set up electromagnetic waves in the æther which spread out radially in all directions from the transmitting aerial, unless the latter has directional transmitting properties. The nature of the propagation process and the phenomena occurring in the medium through which the E.M. wave passes are separately considered in Section " P." It is sufficient here to state that R/F currents in an aerial radiate energy which is propagated through a medium and subsequently arrives at a receiving aerial.

When an E.M. wave encounters an aerial it induces in it alternating voltages and currents of very small magnitude, and of the same frequency as that of the wave itself. The fundamental problem of reception consists in devising apparatus which will make these minute currents perceptible to the senses of sight or hearing. If the amplitude or frequency of the currents in the transmitting aerial can be varied by means of a signalling key (W/T), or by means of the voice (R/T), and these effects can be reproduced at the receiving end, communication will be rendered possible. It will be shown that exact measurement of the currents in the receiving apparatus is not necessary, and that it is only necessary that starts and stoppages and variations of the transmitter currents should give corresponding results in the device which affects the senses, the telephone for aural reception and the undulator (F.62) for automatic high speed visual reception. In this section we are only concerned with aural reception.

2. **The Essentials of a Receiver.**—In order to effect reception of a signal, the receiver must have two essential qualities :—

(*a* It must have the requisite " **selectivity,**" the quality enabling it to select the desired signal to the exclusion of interference from undesired signals.

(*b*) The requisite amplification must be provided in order that the signal may be comfortably audible—" **audibility.**"

Any receiver may be considered, broadly, to consist of **tuning and amplifying circuits,** the object of which is to select and amplify the desired signals, followed by a " **detector** " which turns the H/F currents into the most suitable form for energising the device which will render the signal perceptible to the sense of hearing ; this is usually a pair of **telephones** or loud-speaker.

3. **Necessity for a Detector ; A/F and R/F Components.**—E.M. waves may be produced by R/F currents having frequencies ranging from (say) 16 kc/s. to (say) 10^9 kc/s. Fig. 2 (*a*) represents a simple circuit by which incoming E.M. waves may be made to produce oscillatory currents of

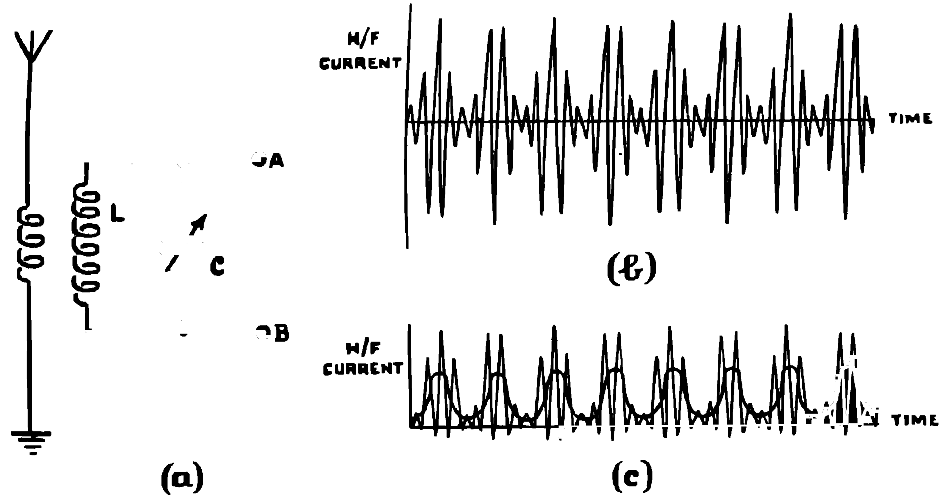


FIG. 2.

the same wave form in the tuned LC circuit ; the incoming signal produces a current in the aerial which injects an E.M.F. into the tuned circuit across the mutual inductive link. The oscillatory current in the tuned circuit would produce an oscillatory voltage across the condenser C, the wave form of which might resemble that of Fig. 2 (*b*). If telephones were connected between terminals A and B, no sound would be heard for the following reasons :—

(*a*) The diaphragm of a telephone earpiece cannot vibrate at frequencies much beyond the audible range, and at radio-frequencies the diaphragm remains unmoved, indicating that **the mean value of the R/F current is zero over any appreciable time in excess of the period of the R/F cycle** ; any two consecutive half cycles of R/F current are approximately equal, and cancel each other out so far as the telephone diaphragm is concerned.

(*b*) Even if the diaphragm could vibrate at radio-frequencies, the ear could not detect the result since it is above the A/F range.

(*c*) The high inductance of the telephones would present such a large impedance to the R/F current that its value would be too much reduced.

In Fig. 2 (*b*) the amplitude of the H/F current is shown to be varying, alternately increasing and decreasing. In this case **it will be assumed that these variations in amplitude succeed each other at a modulating frequency, which is audible,** *i.e.*, between 16 and 20,000 cycles per second. For the reasons given above, no sound can be produced in telephones from this *symmetrical* wave form, principally because its mean value to the insensitive telephone diaphragm, and to the human ear, is zero. To produce an audible result something must be done to render the wave form *asymmetrical*, an effect which can be achieved by cutting off either the upper or lower half of the wave form. Assuming that some device exists by means of which this may be done, the effect of wiping out most of the negative half cycles is the production of a wave form resembling Fig. 2 (*c*). In that case successive half cycles of R/F current do not cancel each other, and **the mean value of the detected H/F current rises and falls at an audio-frequency corresponding with the rate of variation in amplitude of the R/F signal.** In the figure, a full line curve shows the A/F component of the total current, and if this were passed through telephones, vibration of the diaphragm would take place and an audible result would be achieved. The basic function of a **detector** is to produce an asymmetrical variation of applied voltage.

In Fig. 2 (*c*) it will be noted that the H/F current is varying at the same frequency as that of the incoming waves ; but, *in addition*, there is a superimposed audio-frequency change of current. Before detection, the wave is entirely at radio-frequency, but after detection an audio-frequency component is present and may be separated by means of suitable circuits.

The idea of the resolution of an asymmetrical wave form into a symmetrical wave form, the **R/F component,** superimposed on a varying mean value, the **A/F component,** is of great importance in wireless work : Fig. 3 (*a*), (*b*) and (*c*) illustrates this point. The operation of detection may be considered as made up of two parts, the production of an asymmetrical current variation from the input symmetrical voltage wave form, and the analysis by a suitable circuit of this current into its components, of which only the A/F component is necessary for the production of sound in telephones. In general, there is also **a direct current component** resulting from the detection process. (Paragraph 12).

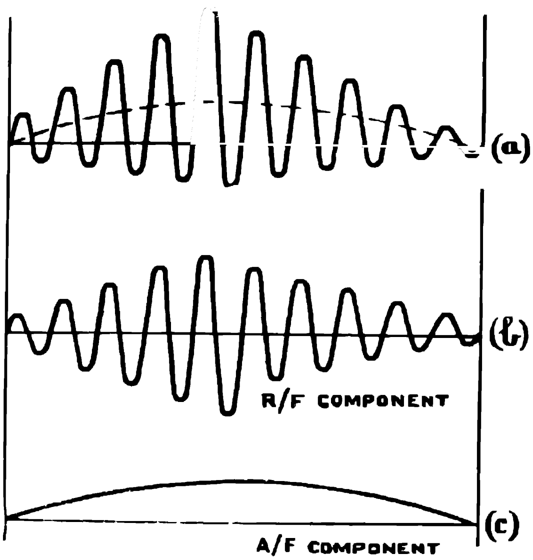


FIG. 3.

4. Production of Audible Sounds Modulation.—From the foregoing work it would appear that an audible sound may be produced after detection, provided that the (*modulating*) frequency with which the E.M. wave passes through its cycle of amplitude variation lies within the audible range. The human ear can distinguish sounds caused by vibrations between the limits of 16 and 20,000 cycles per second, but much prefers frequencies between 800 and 3,000. Most people of normal hearing can hear a bat squeak, but some elderly people cannot ; the average upper limit is about 16,000 cycles per second, which gives a very high pitched squeak. For comparison it may be noted that the middle C on a piano has a pitch of about 250 cycles per second.

Most telephone receivers will also respond to frequencies up to 5,000 cycles, but give best response at frequencies in the neighbourhood of 800-1,000 cycles.

Summarising, in order **to produce an audible sound** it is necessary :—

(*a*) To **modulate the wireless R/F " carrier " wave,** *i.e.*, to break it up into a series of groups or pulses which follow one another at audio-frequency.

(*b*) To **detect the modulated pulses,** the effect of which is to convert each group of R/F waves to one variation of current through the telephone windings, and hence to produce one movement of the diaphragm as the cumulative effect of the many constituent R/F waves in the pulse.

In the spark system, the wave is modulated at the transmitter, because oscillations only occur each time the spark gap breaks down ; the process is represented in Fig. 4, and an audible sound results in telephones because the spark train frequency is within the audible range.

In the I.C.W. system, the wave is modulated at the transmitter by a process which is further discussed in K.52, the wave form resembling Fig. 1 (*c*) or Fig. 2 (*b*). Where the modulation is sinoidal, a pure note may be obtained in telephones, provided that the modulation frequency lies within the audible range ; it should, however, be noted that supersonic modulation has certain

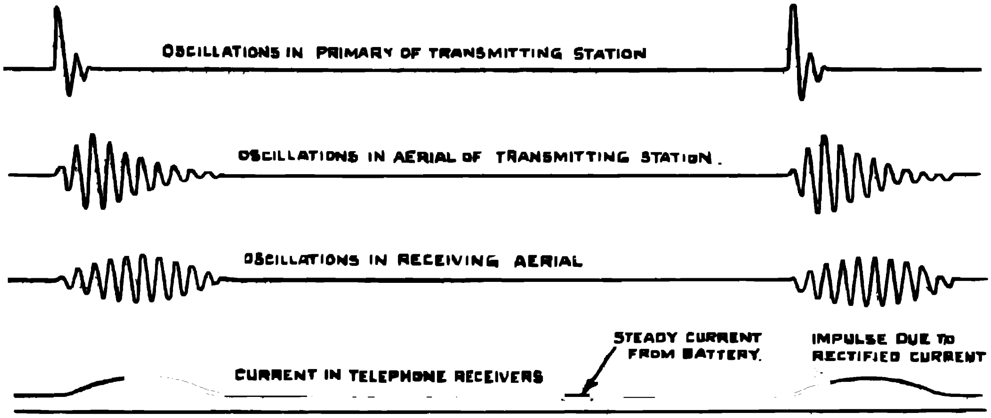


FIG. 4.

uses. For example, using a " quench receiver " (F.56) an audible note may be obtained from S/F modulation ; moreover, for high speed reception (F.62) giving a visual record, supersonic modulation is as good as modulation lying within the audible range.

Expressed somewhat differently, it may be noted that the result of detection is the extraction of a note corresponding to the frequency of variation in amplitude of the *amplitude envelope* of the I.C.W. wave form. The modulating process at the transmitter puts an audio-frequency (or S/F) imprint on a radio-frequency carrier wave ; the E.M. wave travels to the receiver, and the process of detection extracts the A/F imprint, or the S/F one, corresponding to the shape of the amplitude envelope. Essentially, this is also the basis of radio-telephony (R/T) ; the modulation of the I.C.W. wave form may either be sinoidal, *i.e.*, simple in form, or very complex, corresponding to the many different simple sounds composing speech, noise, or music. In this connection it should be noted that the mean (anode) current varies with the amplitude of the incoming signals, but unless this variation is linear, distortion of the modulating signal must result. **For distortionless reception of R/T, linear detection is essential,** but for W/T signals, distortion of the note is not a matter of serious consequence, and usually the mean (anode) current varies with the square of the amplitude of the incoming signal (*cf.* paragraph 16).

In the case of continuous waves (**the C.W. system**), the waves travel from the transmitter devoid of any A/F or S/F imprint. Detection of such waves only results in the production of a click in the telephones, indicating a sudden alteration in mean value of the telephone current. It will be shown, later, that in order to produce an audible result it is necessary to modulate (**heterodyne**) the incoming C.W. waves at the receiver.

5. **Rectification ; Demodulation.**—The process of turning a symmetrical wave form into an asymmetrical one, often done by suppressing one half of it, is given the general name **rectification.** It is, however, fairly common practice to restrict the use of the term rectification to the production of D.C. from A.C. for power supply purposes, using the term **detection** when the object is the extraction of the modulation from an incoming radio signal. Sometimes the word **demodulation** is used instead of detection.

The essential characteristic of a detector is that it should have uni-directional or one-sided conductivity, so that equal amplitudes of voltage applied to it in opposite directions should result in unequal currents passing through the detector. In other words, the resistance of a detector does not conform to Ohm's law ; the detector presents a lower resistance to currents in one direction than in the other.

An incoming modulated R/F signal applies voltage variations to a detector and produces for each complete cycle of voltage variation an excess of current in one definite direction. The nett

result is that after one whole series of R/F oscillations the telephone diaphragms moves out of position once only ; if these series succeed each other at a suitable frequency, as in Fig. 2 (*c*) or Fig. 4, an audible note will be produced

With reference to paragraph 3, it is *usually* necessary to effect a separation of the A/F and R/F components, and the circuit must always have some impedance across which the A/F current may develop a P.D., either for application to an amplifier or directly to actuate telephones.

6. **Series and Parallel Connected Detectors—CR Values.**—Fig. 5 represents two general ways in which a detector may be connected in order to give an A/F output in a load impedance. The " bench mark " symbol may be used to represent any rectifier, and the arrow head indicates that direction of an applied voltage for which it is conducting, and hence also indicates the direction of conventional current through it (*cf.* H.2, 10).

In Fig. 5 (*a*) the condenser C presents a low impedance to the applied R/F signals, and to the R/F component resulting after detection. The resistance R constitutes the **output impedance** for the A/F component, which, in simple cases, may consist of a pair of telephones. In that case it is not always necessary that there should be a separate condenser, the self-capacity of the telephone

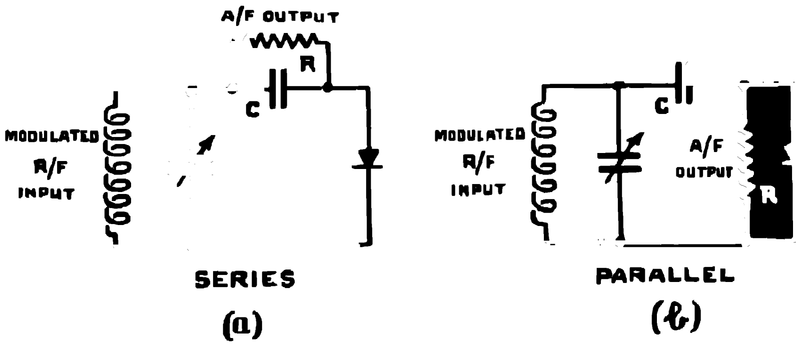


FIG. 5.

windings sometimes sufficing to effect the separation of the R/F and A/F components, and to prevent the building up of R/F potentials across the telephone windings. When C is present, its reactance at the frequency of the R/F component must be small compared both with the load resistance R and the rectifier resistance.

In Fig. 5 (*b*) the load resistance is connected across the detector, and in this case it is *essential* to use a blocking condenser C in the position shown ; if it were not there the A/F component could not build up the requisite P.D. across R, since the latter would be short-circuited by the input circuit.

Both circuits virtually effect a separation of the R/F and A/F components, the condenser in each case providing an easy " by-pass " path for the R/F one ; it provides a path of higher impedance for the A/F one which accordingly chooses the only other path, namely that through R. From another point of view, the effect of the condenser is that it receives unequal positive and negative charges during successive half cycles, and therefore accumulates a resultant charge which sets up a P.D. across the condenser and forces pulses of uni-directional current through the load resistance R, or telephones. In the case of a modulated wave of simple I.C.W. form, these pulses of current will follow each other at the modulating frequency. This point of view is useful, since it brings out the idea that the condenser must be able to discharge itself through the resistance R in time to be ready to receive the next charge from the succeeding R/F signal pulse. This is a matter which depends on the **time constant or CR value** (Vol. I) of the combination, and the latter must be chosen having regard to the modulating frequency in use. The CR value of a detector working well with an I.C.W. signal, modulated by a 1,000 cycle note, would not necessarily give good results in R/T reception, where modulating frequencies up to 8,000 cycles must be considered (*cf.* N.27, 29, 39).

This matter is treated in greater detail below, but it may be stated that the typical CR value in Service detectors is of the order of $2/10^4$ seconds. The latter means that the condenser discharges to one-third of its initial charge in $2/10^4$ seconds, or that the detector device can deal reasonably well with modulating frequencies of the order of $10^4/2$ cycles per second, *i.e.*, 5,000 cycles/second. This is not good enough for the reception of R/T, and a broadcast receiver has a CR value of the order of $1/10^4$, C being typically 0·0001 μF. and R being 1 megohm, a device which will respond adequately to modulating frequencies of the order of 10,000 cycles per second.

In other words, the time constant should be large in comparison with the period of the R/F component, but small in comparison with the A/F one.

7. **A Simple Receiver having a Coupled Secondary Circuit.**—Fig. 6 represents a simple form of receiving circuit which may be used in a preliminary discussion of broad principles. It may be assumed, if desired, that a crystal detector is being employed.

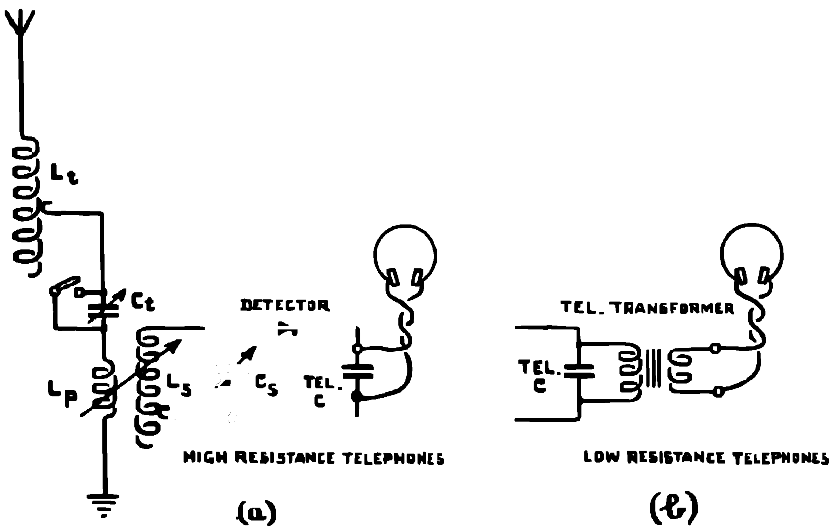


FIG. 6.

Since an E.M. wave induces in a receiving aerial E.M.Fs of the same frequency as itself, the resulting oscillatory current in the aerial will have a maximum value if the aerial is tuned to the same LC value as that which corresponds to the frequency of the wave, *i.e.*, the same LC value as that of the transmitting aerial. The greater the oscillatory current in the receiving aerial, the greater will be the oscillatory voltage which can be applied to the detector, and the greater will be the audibility of the signal. In some cases, therefore, the aerial circuit is tuned, or made resonant to the voltage induced in it. We may regard the effect of the incoming wave as being represented by an alternator inserted in the aerial circuit, giving an R/F voltage of small amplitude which produces forced oscillations in the aerial.

The aerial circuit is shown consisting of an **aerial tuning inductance** L_t, and an **aerial tuning condenser** C_t, for adding capacity in series with the aerial. The circuit is completed through the aerial capacity σ and the resonant current flows through L_p, the primary of a mutual inductive coupling to the secondary circuit.

The detector may be connected up across the inductance or condenser of the aerial circuit itself, but such an arrangement is very unselective. In general, a detector, or receiver, is fed from a separate circuit of its own, inductively or otherwise coupled to an aerial circuit. The secondary circuit is tuned to have the same LC value as the aerial circuit. Resonant currents will therefore circulate in it, and the oscillatory voltage set up across the secondary condenser by these currents is applied to the detector. It is shown later that in order to apply the greatest possible voltage to

the detector, L_s should be made large and C_s should be made small, since the value of the voltage across the condenser is given by $I_s\sqrt{\frac{L_s}{C_s}}$, where I_s is the secondary current. This argument neglects the effect of any damping due to the detector, a serious factor varying with the input.

In this way, R/F oscillatory voltages of as large amplitude as possible are applied to a detecting device, following which the R/F component passes through the condenser, usually known as **the telephone condenser,** and the A/F component passes through the telephones. In cases where the telephones do not present the requisite high impedance to the A/F component, connection is made by means of a step-down **telephone transformer,** the latter effectively stepping up the resistance of the telephones to the requisite value (Vol. I). The use of a telephone transformer also protects the telephones from the passage of any D.C. component which may result from a rectification process, and whose presence may be undesirable.

8. **Numerical Example on Aerial Tuning.**—It is required to receive (*a*) a wave whose LC value is 400 ; (*b*) a wave whose LC value is 20, on an aerial whose natural inductance is 30 mics., and natural capacity σ is 1·5 jars. The primary (L_p) of the inductive coupling to the secondary circuit has a value of 40 mics.

(*a*) Total inductance required $= \frac{LC}{\sigma} = \frac{400}{1\cdot5} = 266\cdot6$ mics.

Inductance already in circuit $= 30 + 40 = 70$ mics.

Additional inductance required on tuner $= 196\cdot6$ mics.

(*b*) Total capacity required $= \frac{LC}{L} = \frac{20}{70} = \frac{2}{7}$ jar.

Capacity already in circuit $= 1\cdot5$ jars.

Additional capacity (C_t) must be added such that

$$\frac{1}{C_t} = \frac{1}{C} - \frac{1}{\sigma} = \frac{7}{2} - \frac{1}{1\cdot5} = \frac{7}{2} - \frac{2}{3} = \frac{17}{6}.$$

$$C_t = \frac{6}{17} = 0\cdot35 \text{ jar.}$$

9. **Types of Detector.**—In order to detect a signal, many different devices have been used. The most important, in historical order, are the coherer, the electrolytic detector, the magnetic detector, the crystal detector, and the valve detector. There is also the metal rectifier, a device which is separately described in H.10 ; it may be regarded as the modern form of crystal detector, but is used chiefly for power rectification.

These differ widely in their appearance and nature, but all achieve the same purpose, *i.e.*, to turn a group of R/F oscillations into a single current variation, and hence a number of groups themselves recurring at audio-frequency into an A/F current variation, through the device which renders signals perceptible to the ear.

In present day practice, both metal rectifiers and valves find application as detectors, the latter being much more commonly used than the former.

As, however, the crystal detector illustrates very conveniently the principle of rectification, without the complication produced in a receiving circuit by the introduction of a valve, a brief description of its action will now be given.

10. **The Crystal Detector.**—We have seen that the essential feature of a detector is that its conductance should vary with the direction of the applied voltage.

A peculiar property of certain combinations of two crystalline substances in contact with each other, or of a crystal in contact with a metal, is that they form a conducting path whose resistance varies according to the direction and also the amplitude of the voltage applied across them. In

other words, (*a*) for equal applied voltages they allow more current to pass in one direction than the other, and (*b*) the ratio of voltage to current for varying voltages in the same direction, is not a constant ; Ohm's law is not obeyed.

The actual values of resistance, for crystals which were in common use, is high, and of the order of 10,000 to 100,000 ohms. For small applied voltages the current may even be zero, and the resistance infinite. Carborundum and steel is a crystal-metal combination often found in earlier days. A fragment of the crystal was mounted in a small brass cup and the metal contact or " cat's whisker " was held against it by a spring. The sensitivity of the combination depends very much on the contact pressure, and the sensitivity of different " spots " varies.

The theory of the action of crystal couples, or crystal metal combinations, cannot be said yet to be thoroughly understood.

11. **Characteristic Curves.**—The action of a detector device, such as a crystal, can be appreciated by considering its " characteristic curve," which is a graph of current against voltage for varying values of voltage in both directions.

Fig. 7 (*a*) represents an ordinary resistance in series with an ammeter between the centre point C of a potentiometer and the slider E. In this way, equal voltages in opposite directions may be applied to the resistance, the resulting current being measured by the ammeter.

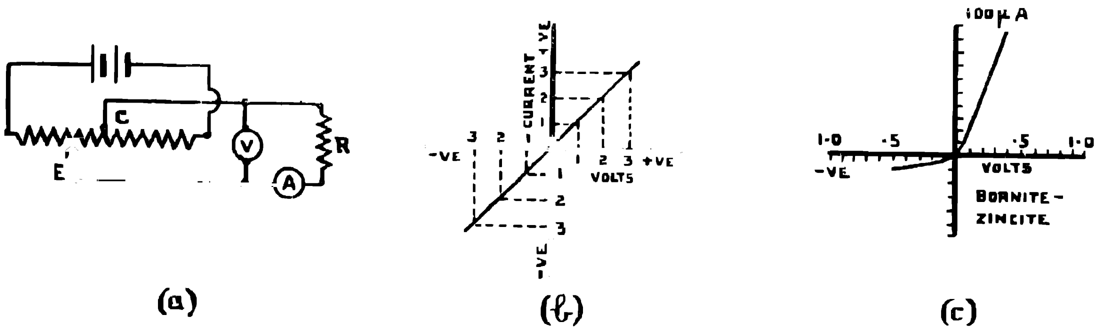


FIG. 7.

Fig. 7 (*b*) shows the result of plotting current against voltage ; this curve is called a characteristic curve of the resistance. From it we see that when C is 1-volt positive to E a current of 1 ampere flows from C to E through the resistance, for 2-volts, 2 amperes flow, and so on. Similarly, when C is 1-volt negative to E, a current 1 ampere flows in the reverse direction and so on.

From this we gather that the resistance R is given by

$$R = \frac{V}{I} = \frac{1}{1} = \frac{2}{2} = 1 \text{ ohm.}$$

If the resistance R is replaced by a detector device, such as a crystal detector, we get a totally different characteristic, the form of which depends on the nature of the device. Fig. 7 (*c*) represents the characteristic curve obtained by the bornite-zincite combination, in which the direction from bornite to a zincite is reckoned " positive."

The characteristic curve indicates that when a voltage is applied in the positive direction, the resistance of the detector is low, *i.e.*, 0·1 volt gives a current of 20 microamperes, 0·3 volt gives 72 microamperes, etc. When voltage is applied in a negative direction the resistance of the device increases enormously ; —0·1 volt gives about 3 microamperes, and —0·5 volt only gives about 12 microamperes.

Hence the resistance of the detector to voltages applied in a positive direction is much less than its resistance to voltages applied in a negative direction.

Consider now what happens when an incoming R/F signal applies an oscillatory voltage across the detector in the manner shown in Fig. 6. The process is represented in Fig. 8 (*a*), which shows the signal applying voltages alternately in opposite directions across the detector device, in this case the crystal.

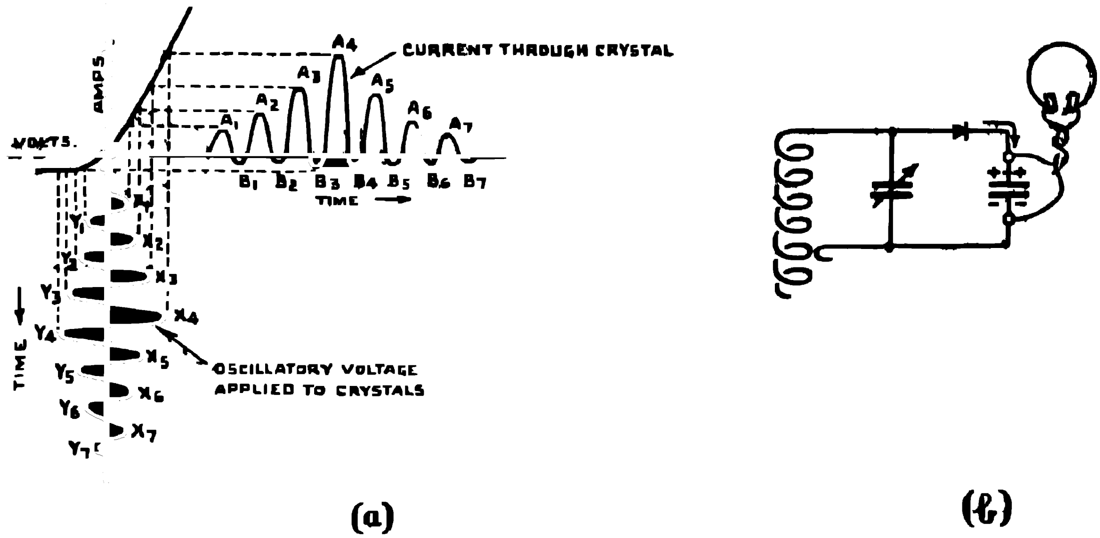


FIG. 8.

It is important to notice the way in which the figure is drawn. By running perpendiculars from the oscillatory *voltage curve* to cut the characteristic curve, the corresponding *current curve* can be determined. The axes of co-ordinates are made to serve both as axes of current and voltage, and as axes of time.

From the figure it is clear that the first result of the introduction of the detector is to produce an asymmetrical variation of current from a symmetrical variation of applied voltage, the frequency with which the current varies being the same as the frequency of the incoming signal. We have therefore achieved the result which was shown to be desirable in paragraph 3; the mean value of the current now alternates at a frequency depending on the modulation, and will affect the telephones if the modulation frequency is within the A/F range.

Fig. 8 (*b*) shows the portion of Fig. 6 which includes the detector. As a result of the process shown in Fig. 8 (*a*), the condenser receives unequal positive and negative charges during each successive half cycle and, therefore, accumulates a resultant charge which sets up a P.D. across the condenser, making the diaphragm move out of position once only as a result of the passage of this part of the wave train. Alternatively, as in paragraph 6, it may be considered that the object of the condenser is to separate the two components resulting from detection by providing two circuits in parallel, the condenser for the R/F component and the telephones for the A/F one.

The basic process which has been outlined above is true of any detector.

12. **The Diode Detector.**—Valve detectors have almost completely replaced all other forms, and the simplest of these is the diode. As explained in Section "B," electrons only flow appreciably from filament to anode of the diode valve, when the anode is positive with respect to filament or cathode. From this point of view it is clear that the diode forms an almost ideal one-way device or "valve," and its application as a detector is clearly indicated.

Fig. 9 (*a*) represents a series connected diode detector, Fig. 9 (*c*) representing the parallel connected circuit (paragraph 6). In each case, R is the output impedance across which potentials at the modulating frequency are developed, and C is the R/F by-pass condenser ; in the simplest case,

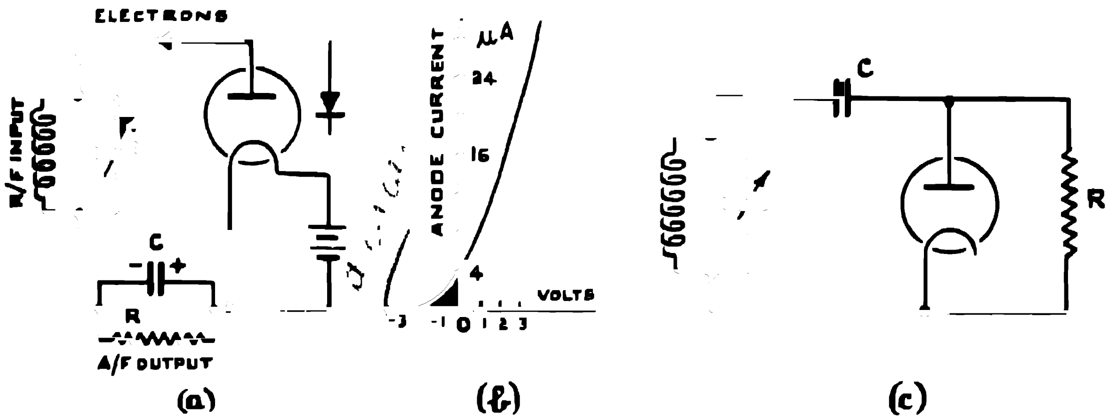


FIG. 9

the resistance R might be replaced by telephones, but it is more usual to pass on the rectified A/F output for subsequent amplification. The bench mark symbol beside Fig. 9 (*a*) indicates the direction in which conventional current flows across the valve ; with thermionic detector devices it is always more convenient to think in terms of electrons, which flow in the opposite direction. The side of the condenser attached to the filament acquires a positive charge, and the other side acquires a negative charge due to the accumulation of electrons on it. If the resistance R were not connected across the condenser in Fig. 9 (*a*), electrons would continue to accumulate on the side of the condenser attached to the anode, until the P.D. across the condenser became equal to the peak voltage of the incoming R/F signal. When that condition is reached no further electrons will flow across the valve ; this is the principle of operation of the peak voltmeter (H.2). By connecting resistance R across the condenser the voltage across the latter falls slightly from the peak value and, in the case of an unmodulated signal (C.W.) **a steady D.C. current** flows through R (*Cf.* Fig. 24 in N.27). In the case of I.C.W., a current at the modulating frequency also flows through R, being superimposed upon the " **D.C. component.**"

Fig. 9 (*b*) represents approximately the characteristic curve of a diode. It will be seen to resemble that of Fig. 8 (*a*), and the mechanism of the detection process is essentially similar to that case.

The diode detector is more fully described in Section N.27 ; it was the first and simplest of the valve detector devices, but its comparative insensitivity led to the development of other forms of valve detector. It is now being used much more than when it was first discovered, especially in superheterodyne receivers where the detector often follows stages of amplification and extreme sensitivity is not required. For R/T work it has the special advantage that its characteristic is very straight, a matter of importance when it is desired to avoid distorting the modulating signal ; the same feature gives it a special advantage for use in A.V.C. systems (N.35).

13. **The Double Diode.**—With the circuit of Fig. 9, electrons only appreciably flow across a valve when the anode is positive with respect to the cathode, *i.e.*, during the positive half cycles of input voltage. By using a double diode and a centre tapped input circuit, it is possible to have electrons flowing across the valve during each half cycle

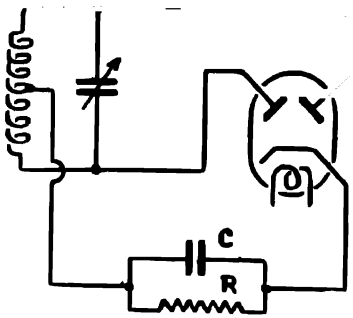


FIG. 10.

Fig. 10 represents a double diode in a series connected circuit. If there were no load resistance R connected across the condenser, electrons would accumulate until the P.D. across the condenser is equal to the peak value of the input signal, after which the action would cease.

This circuit is similar to that of the full wave rectifier used in the production of D.C. from A.C. for power supply purposes (H.3). In this simple form it is seldom used in R/F circuits, being more usually found in connection with the valve known as the double-diode-triode (N.38, B.45), a valve in which one diode detects, the other provides automatic grid bias, and the triode portion amplifies the detected signals.

14. **The Triode as a Detector.**—Since the essential feature of a detector is the possession of a non-linear characteristic, the work of Section " B " makes it evident that the triode may be used by taking advantage of either—

(*a*) the non-linear characteristic curve of anode current plotted against grid voltage, utilised at either its upper or lower bend where the curvature is variable. This is termed **anode bend rectification,** or

(*b*) the non-linear characteristic curve of grid current plotted against grid voltage, using the grid-filament circuit of a triode as if it were a diode detector, subsequently applying the A/F signals resulting from detection between grid and filament of the triode portion of the valve, the latter acting as an amplifier. This is termed either **cumulative grid rectification,** or **grid current rectification.**

15. **Anode Bend Detection.**—Fig. 11 (*a*) represents the circuit details of a simple anode bend detector, including a potentiometer by means of which the steady potential of the grid relative to the filament can be adjusted, in order to set the working point to the middle of the upper or lower

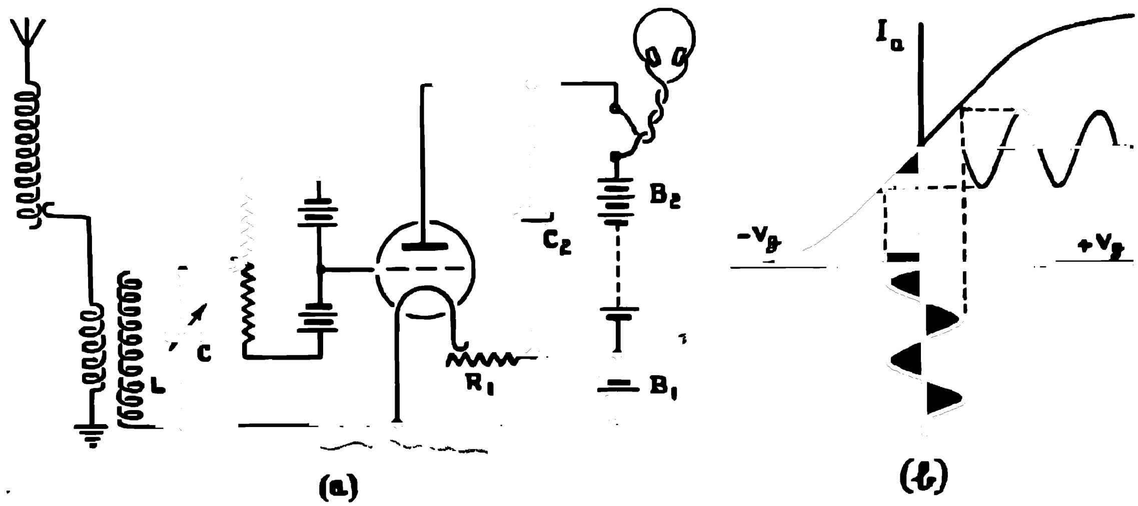


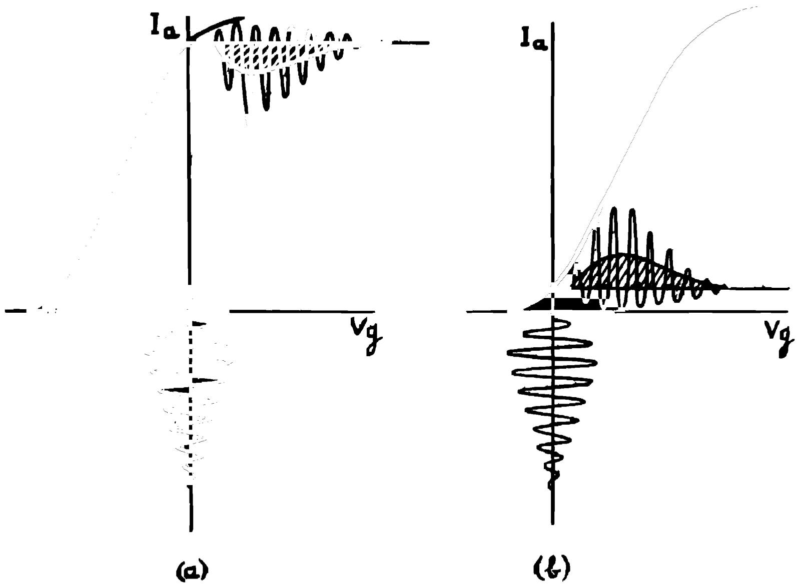
FIG. 11.

bend. The signal applies an input between grid and filament symmetrically with respect to the working point, lower anode bend detection producing a result very similar to that of Fig. 8 (*a*).

The output impedance consists of a pair of high resistance telephones for the A/F component, and the relatively large condenser C_2 for the R/F one.

It may be noted, from Fig. 11 (*b*), that no detection takes place if the grid bias is adjusted so that the working point falls on the straight portion of the characteristic; the mean anode current is unaltered by the signal.

Fig. 12 (*a*) and (*b*) represents the process of upper anode bend and lower anode bend detection respectively, in the case of an I.C W. signal.



ILLUSTRATING ANODE RECTIFICATION.

FIG. 12.

Before the signal arrives, in either case, a steady anode current will flow through the anode circuit of an amount given by the length of the ordinate.

On receipt of a modulated signal, using **upper bend detection**, the current decreases more during the negative half cycles of oscillatory grid voltage than it increases during the positive half cycles. The nett result is that the **mean anode current shows a decrease** during the time occupied by one A/F modulating cycle. A heavy D.C. current is drawn from the H.T. supply, and flows through the telephones.

In the case of **lower anode bend detection** the opposite effect occurs, the **mean anode current showing an increase** over the same period of time.

Lower anode bend detection is the form commonly used. It has the advantage that the normal adjustment makes the grid negative to the filament, and no grid current flows. The effect of grid current flowing, is that energy is absorbed from the input circuit and its damping is increased (paragraph 23). This may be contrasted with "grid current detection," in which the flow of grid current is an essential condition. A further advantage of lower bend detection is that less current is taken from the H.T. supply. Moreover, the anode bend detector is considerably more sensitive than the diode or the crystal, because of the amplifying properties of the triode.

16. **Anode Bend Detection of Weak Signals.**—The performance of an anode bend detector depends very much on the amplitude of the incoming signals, and it is possible to distinguish two special cases, "**power anode detection**" and "**weak signal detection**" respectively.

Power anode detection involves the application of strong signals to the detector, the latter giving an output varying approximately in a linear manner with the input; this is a matter of considerable importance for the reception of R/T, where it is essential that the operation of detection should not distort the wave form of the amplitude envelope of the modulated signal. It gives an approximation to the ideal **linear detector** and is separately treated in Section N.26, 29; for this reason R/T signals need amplification before power anode detection.

For W/T purposes, one is mostly concerned with the reception of weak signals, and it is not a matter of considerable importance if the note corresponding to the modulation of the wave form is slightly distorted. With **weak signals** it can be shown that the output current is approximately proportional to the square of the input oscillatory voltage. This "**square law detection**" follows directly from the use of the curved portions of the mutual characteristic, since a small portion ofany curve may be represented approximately by a square law of parabolic form; a brief mathematical treatment is given below (paragraph 18).

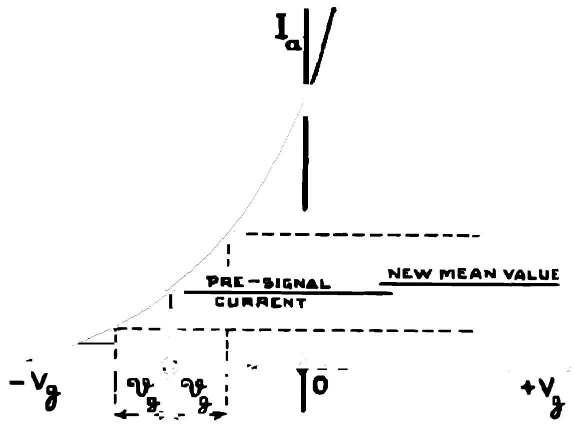


FIG. 13.

Fig. 13 shows clearly the dependence of the detector device on the curvature of the characteristic [*cf.* Fig. 11 (*b*)]. Under no-signal conditions, the operating point is to the left of the origin O, in the region of negative grid bias. There is a steady pre-signal current, indicated in the figure. On receipt of an unmodulated C.W. signal of peak value $\mathcal{V}_g$, the anode current varies between the limits shown and **the mean value rises** to the line marked "new mean value." The only response given by the telephones during a group of C.W. oscillations representing a morse dot or dash is, therefore, a click at the beginning when the diaphragm moves from one steady position to another, and similarly at the end of the signal. This would not be a satisfactory way in which to receive a C.W. signal, and it is shown later that a note is impressed on to it by modulating it (*heterodyning*) at the receiver.

In the case of I.C W. signals of small amplitude, the mean value of anode current during the receipt of a signal will vary in accordance with the modulation on the wave; the anode current does not, however, vary in a *linear* manner with the amplitude of the input signal.

Fig. 22 in Section "N" represents the operation of anode bend detection using input signals of large amplitudes; the study of linear detection may conveniently be deferred until reading that section.

17. **Adjustments for Anode Bend Detection—Characteristic Used.**—The necessary adjustment for good detection is that the normal grid voltage should correspond to a point on the characteristic at either the upper or lower bend.

In the case of **upper bend detection** this may be obtained by :—

(*a*) Filament control.—Alteration of heating current to the filament alters the point on the I_a/V_g curve where it approaches saturation. In practice, the filament is dulled by increasing the resistance in its heating circuit.

(*b*) Alteration of anode potential.—This shifts the characteristic to the left when increased and to the right when diminished.

(*c*) Alteration of grid potential by potentiometer control.

In the case of **lower bend detection** the following methods only are applicable :—

(*a*) Alteration of anode potential.—This shifts the characteristic to the left when increased, and to the right when diminished.

(*b*) Alteration of grid potential by potentiometer control.

On account of the impedance in the anode circuit, the dynamic characteristic is, effectively, the one used. Hence, the characteristic is flatter than in the static case, and, when using lower bend detection, the difference in slope of the curve on the two sides of the bend will not be so great as if it were the static characteristic.

The efficiency of the circuit as a detector for weak signals is greater the greater the curvature, or the rate of change of slope at the rectifying point, as will be shown mathematically below. Hence, a valve should be used for anode bend detection whose A.C. resistance is high compared with the external impedance, so that the slope of the dynamic characteristic does not differ greatly from that of the static. In addition, the steeper the slope of the actual characteristic, the sharper will be the bend of the curve. Hence, a detector valve should have a high value of r_a and a high value of g_m, and in consequence a high value of m. This gives also the advantage that the characteristic falls to zero at a voltage which is not too negative, and so reduces the amount of grid bias necessary.

★18. **Mathematical Note on the Square Law Detection of Weak Signals.**—In general, the value of the rectified current depends in a complicated manner on the amplitude of the applied signal voltage, and no simple relation between the two can be stated, except in the case of signals of small amplitude. The analysis shown below only applies for small input voltages to the detector. In modern receivers, with a number of R/F amplifying stages, the practical results may be very different.

For a small part of its length, round a point where its slope is changing, a curved characteristic can be taken as approximating to a parabola, and so can be represented by ane quation of the form $I_a = a + bV_g + cV_g^2$ where a, b and c have certain definite values depending on the curvature of the dynamic characteristic in the region.

Let the polarising, or steady, voltage applied to the grid before a signal arrives be V_0. Then a steady anode current I_0 flows, given by $I_0 = a + bV_0 + cV_0^2$.

Let an oscillatory signal voltage, $V_s \sin \omega t$, be applied to the grid.

This, being superimposed on the steady voltage V_0, gives a grid potential $V_0 + V_s \sin \omega t$. The corresponding current is then

$$\begin{aligned} I_a &= a + b(V_0 + V_s \sin \omega t) + c(V_0 + V_s \sin \omega t)^2 \\ &= a + bV_0 + cV_0^2 + bV_s \sin \omega t + 2cV_0V_s \sin \omega t + cV_s^2 \sin^2 \omega t \\ &= I_0 + bV_s \sin \omega t + 2cV_0V_s \sin \omega t + \frac{cV_s^2}{2}(1 - \cos 2\omega t) \\ &= I_0 + \frac{cV_s^2}{2} + (bV_s + 2cV_0V_s)\sin \omega t - \frac{cV_s^2}{2}\cos 2\omega t. \end{aligned}$$

There are two radio frequency terms in the above expression, whose mean value over a complete cycle or number of cycles is zero. In other words, they are by-passed by the telephone condenser. The term $\frac{cV_s^2}{2}$ represents the change in the **mean** value of current passing through the telephones (*cf.* Fig. 13). Now

$$I_a = a + bV_g + cV_g^2,$$

$$\therefore \quad \frac{dI_a}{dV_g} = b + 2cV_g,$$

$$\text{and} \quad \frac{d^2I_a}{dV_g^2} = 2c.$$

Hence the increase in the mean anode current, $\frac{cV_s^2}{2}$, may be written as $\frac{V_s^2}{4} \times \frac{d^2I_a}{dV_g^2}$.

This shows that **the rectified current is proportional to the square of the amplitude of the applied voltage,** and also to the second differential coefficient of I_a with respect to V_g, *i.e.* to the rate of change of slope of the (I_a, V_g) curve. The most sensitive position for rectification is thus where the slope is changing most rapidly, *i.e.*, at the upper and lower bends of the curve.

It also follows from this result that, with weak signals, it is advisable to amplify before detection, because of this " square law " which connects the final current change through the telephones, and the strength of audible signal, with the value of applied voltage ; if the input signal is doubled, the A/F output will be quadrupled.

19. **The Grid Current Detector—Cumulative Grid Detection.**—Fig. 14 (*a*) and (*b*) represents the series and parallel connected form of a circuit using the non-linear characteristic curve of grid current plotted against grid voltage (paragraph 14).

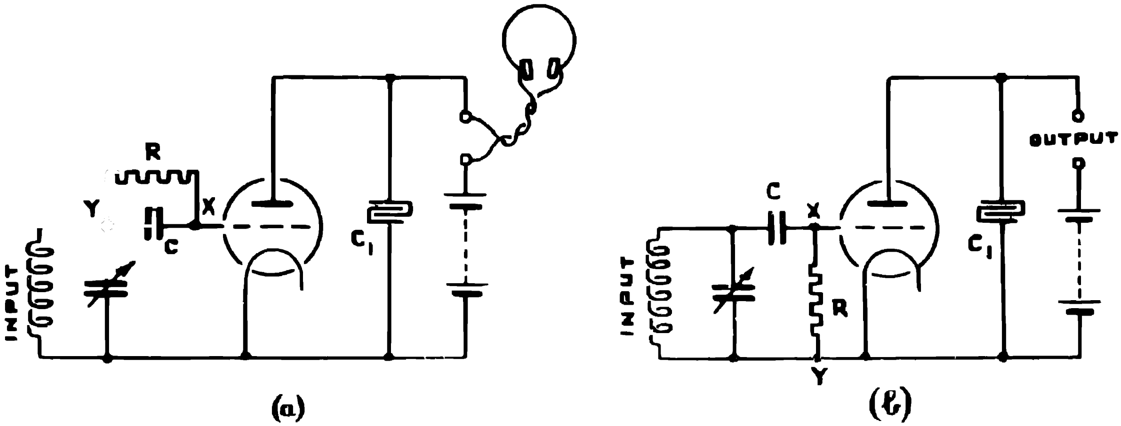


FIG. 14.

Both circuits resemble Fig. 11 (*a*), but have in addition the " **grid leak** " and condenser combination R and C respectively. Moreover, this addition makes the circuit between grid and filament strikingly similar to Fig. 9, the circuits for the diode detector. The latter, in fact, provides the clue to the operation of the grid current detector, sometimes called the " leaky grid " detector.

In Fig. 15 (*a*), the dotted curve represents the incoming I.C.W. signal voltage which is applied between the grid and filament of the valve. As in the case of the diode, electrons flow (**grid**

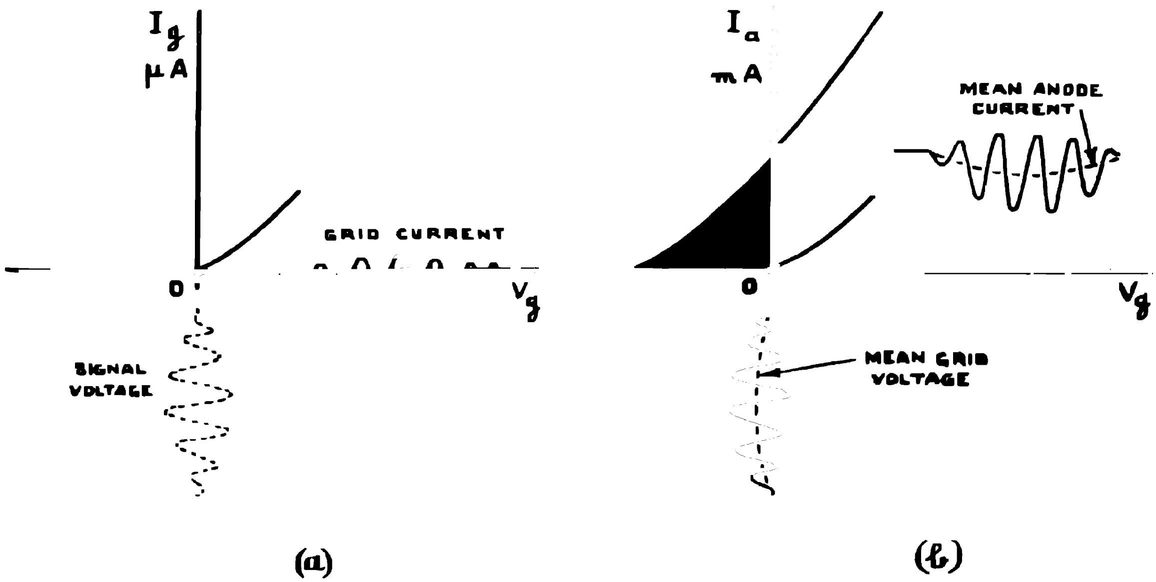


FIG. 15.

current) during the positive half cycles and, if there were no **grid leak** R, a P.D. would be developed across the plates of the condenser equal to the peak value of the incoming signal voltage. With the grid leak R in position, the *D.C. component* (paragraph 12) flows in R, having superimposed upon it the *A/F component* corresponding to the modulating frequency.

Fig. 15 (a) shows the grid current/grid voltage characteristic, and shows the small pulses of rectified grid current which flow during the positive half cycles. As the amplitude of the signal increases, a bigger negative charge is built up on the side of the condenser attached to the grid, and the value of the mean anode current falls correspondingly. The P.D. developed across R, between the points X and Y, is applied between grid and filament of the valve considered as a triode. The latter acts as a class "A" amplifier (N.23, F.8), working on the straight portion of its characteristic, and gives an *amplified* anode current output, the shape being a replica of the input signal. Fig. 15 (*b*) represents the mutual characteristic of the valve, and shows the symmetrical signal input superimposed on a varying mean grid voltage, producing, in consequence, a similar variation of mean anode current, the anode current decreasing periodically at the modulating frequency.

The anode current varies at both radio and audio-frequency, the R/F component being by-passed by the relatively large condenser C_1. It should be noted that the flow of grid current is an essential condition to the operation of the detector. With some valves, grid current does not start until the grid is slightly positive to the negative end of the filament. For this reason it is *sometimes* the custom to connect the end Y of the grid leak to the positive terminal of the filament supply ; a small positive bias will thus be applied which will ensure efficient operation.

It may be shown that, as in anode bend detection, the strength of signals in the telephones is proportional to the square of the amplitude of the incoming signal voltage for small inputs (**square law detection**). The formula is involved, and depends, as might be expected, on the slope and rate of change of slope of the grid current characteristic, as well as on that of the anode current curve. For this reason the grid current detector will distort an R/T signal (paragraph 4), a matter of no consequence in the reception of W/T. In the case of R/T there is a further possibility of distortion if the incoming signal has too large an amplitude. The reason for this is evident from Fig. 15 (*b*) ; if the A/F signals are too large, the mean grid voltage will swing the input to the neighbourhood of the lower bend of the mutual characteristic, and lower bend detection will take place. Since the latter tends to *increase* the mean anode current, it runs counter to grid current detection, which tends to *decrease* it, and distortion is introduced.

The treatment of large input signals is considered under the heading "**power grid detection**" (*cf.* paragraph 16) in Section N.30.

Incoming C.W. oscillations cause the grid to take up a state of equilibrium at some potential more negative than its original value ; the electrons flowing into the condenser during the positive half cycle of grid/filament voltage are just equal to the total discharge of electrons through the *leak* during the whole of each cycle. [*Cf.* Fig. 24 (*a*), Section "N."]. The mean anode current *falls* from its pre-signal value, and, again, a click is heard in the telephones.

From the foregoing work, it appears that the grid current detector acts as a triode amplifier following a diode detector (*Cf.* N.27, Fig. 25). For comparison, the anode bend detector may be regarded as an amplifier followed by a detector. For weak signals the grid current detector is the more sensitive of the two.

20. **CR Value of the Grid Leak and Condenser Combination.**—The determination of the most suitable values for C and R is a matter of some difficulty. For audibility, the amplitude of the A/F component resulting from the detection process should be large, a matter which implies that sufficient asymmetry should be introduced into the rectified current waveform ; the mean value of the grid potential must go sufficiently negative. It is also essential that the charge should leak away during the later portion of the A/F cycle. Such considerations give the following results :—

(*a*) C should be small, so that small accumulations of electrons may charge it up to a considerable potential.

(*b*) The product CR should be adjusted to give the requisite "leak away" during the portion of the cycle available. The importance of the "time constant" CR has already been stressed in paragraph 6.

(*c*) The resistance R should be such that its value is much higher than the reactance of the condenser to the signal voltage oscillations; the R/F voltage applied between grid and filament will then be as large as possible, and R will be a suitable impedance for the A/F component. This indicates a high value of R and a high value of C. In any case, R must not be too low, or the rectified charge will drain away practically as fast as it collects.

Since considerations (*a*) and (*c*) give opposing results as regards C, its value must be a compromise, and the value of R is chosen to satisfy (*b*) as well as possible.

21. **Adjustment and Mal-adjustment of the Grid Current Detector.**—With cumulative grid detection, the audio frequency decrease in anode current through the telephones is greater the greater the slope of the anode characteristic. So far we have only considered the action as taking place with **the working point on the straight part of the curve.**

Working point too low down.—Let us suppose that the anode current curve is such that, instead of the straight steep portion of the curve, the lower curved portion is opposite the asymmetrical grid voltage variations. In this case there is less response in the telephones, for two reasons.

Firstly, the slope of the characteristic is less, and, secondly, there is an anode rectification effect due to the concavity upwards of the characteristic. Now cumulative grid rectification always gives an audio frequency **decrease** in anode current, while lower bend anode rectification gives an **increase**, and hence the two effects are in opposition.

Adjustment.—To move the working point further up the characteristic, the anode voltage may be increased, or the grid leak may be connected to the positive instead of the negative side of the filament. This last method may be explained by the fact that, in cumulative grid rectification, with a leak fitted, the initial negative voltage on the grid is not exactly that at which no grid current flows (as we saw it was with the condenser only), but some other voltage determined by the value of the leak itself.

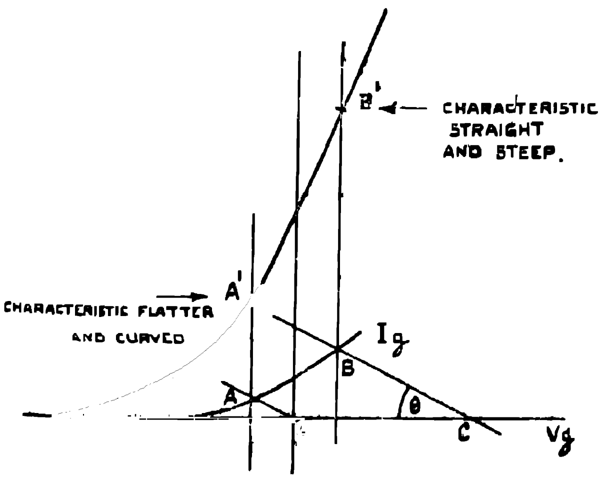


FIG. 16.

The steady voltage on the grid, with no incoming signal, is actually determined by the intersection of the (I_g, V_g) curve and the line OA drawn from zero grid volts (when the grid is connected to the negative side of the filament), in such a direction that its slope is given by $\cot\theta = R$, where R is the value of the leak resistance. In other words, the potential of the grid with respect to the filament is given by $V_g = -I_g R$. In addition, V_g and I_g are connected by the relationship expressed by their characteristic curve, and with both conditions to be satisfied, the construction of Fig. 16 determines the initial potential of the grid (*cf.* N.27).

In the figure shown, the point of intersection A is opposite to A′ on the (I_a, V_g) curve, which is rather low down. If, however, the grid were connected to the positive end of the filament, OC being the voltage drop along the filament, then by the same argument, when angle BCO = θ, the steady voltage on the grid is B, opposite a point B′ on the (I_a, V_g) curve, where the latter is steep and straight.

Working point opposite the top bend of the (I_a, V_g) curve.—Since top bend anode rectification, which is bound to occur in this case as well as cumulative grid rectification, gives a **decrease in** mean anode current through the telephones, the effects are here **additive,** instead of being in opposition, as in the last case.

It is not usually possible to achieve this combination of results because :—

(i) Cumulative grid rectification depends on grid current, and it is usual to employ low values of anode voltage to give bigger values of grid current. This means that the (I_a, V_g) characteristic is moved to the right, and so the upper bend tends to be away from the region of operations.

(ii) If it is proposed to overcome the above difficulty by reducing the filament current, so as to bring the upper bend of the (I_a, V_g) curve to the left, similar difficulties arise. To get good values of grid current a large filament heating current is necessary, and this increases the saturation current and moves the upper bend to the right.

Both from the point of view of filament current and anode voltage, therefore, the conditions which are suitable for upper bend anode rectification are antagonistic to the production of the values of grid current necessary for cumulative grid rectification.

22. **The Moullin Voltmeter.**—The principle of the anode bend detector may be usefully applied to the measurement of R.M.S. voltages at high frequencies. Fig. 17 (*a*) represents the circuit details of one form of Moullin voltmeter, which is essentially a lower anode bend power detector (paragraph 16), in which the increase in mean anode current, consequent upon the detection

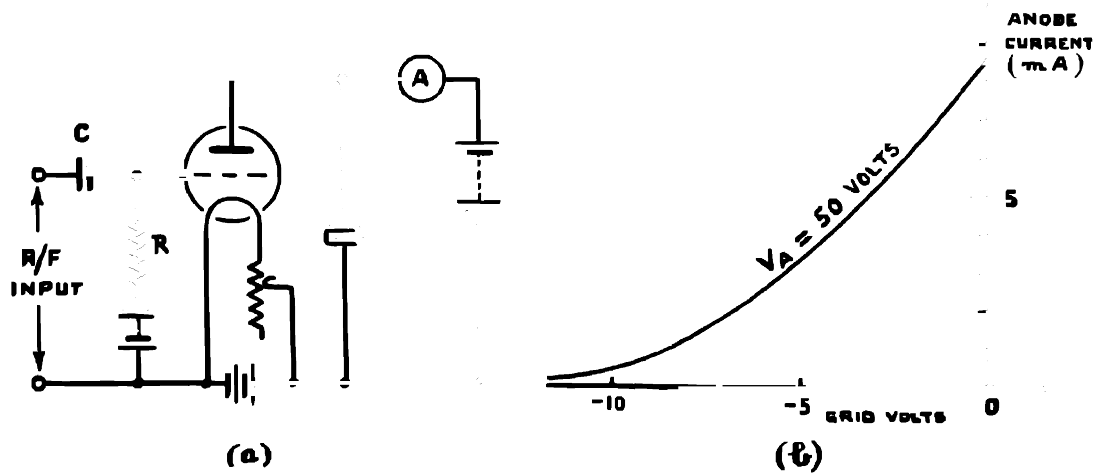


FIG. 17.

of a signal, is recorded by a milliammeter ; the latter may be calibrated to read directly in volts. The variation in anode current is approximately proportional to the input voltage (*cf.* Fig. 20), but is almost independent of the frequency of the applied E.M.F. ; for this reason the voltmeter may be calibrated using a low frequency supply.

These voltmeters are usually low reading instruments, and the range and values of the components will depend upon the characteristics of the valve. For example, a valve having the characteristics shown in Fig. 17 (*b*) could be used to give satisfactory results for input voltages of the order of 5 volts. It would be necessary to make provision for a negative grid bias up to about 10 volts, using an H T. battery of 50 volts ; C and R might have the values of 1 μF and 3,000 ohms respectively.

This type of instrum[illegible] be used to measure the A.C. voltages in circuits also carrying D.C. voltage; the blocking condenser C prevents the applications of the latter between grid and filament of the instrument. In commercial forms of this voltmeter it is usually necessary to adjust the grid bias in order to bring the pointer to the zero of calibration. The maximum input to the instrument is limited by the voltage at which the positive peaks just run into the region in which grid current flows (*cf.* Fig. 20).

The anode bend detector receives a further application in **wavemeters** as a **resonance indicating device** (*cf.* W.13). It might also be used as a monitoring control, indicating the varying "depth of modulation" in the case of an R/T transmitter (*cf.* N.14).

Voltmeters of this type are calibrated using pure C.W. oscillations; the errors introduced by other wave forms may be large.

★23. **Damping of the Grid Current and Diode Detectors.** —The cumulative grid detector (and **the diode**) suffers from the serious disadvantage that it produces a damping effect on the tuned grid circuit, being equivalent to the introduction of a small series resistance. The necessary grid current reduces the **effective parallel resistance** of the grid-filament path from almost infinity, when no grid current flows, to some much lower finite value under working conditions; the lower anode bend detector suffers no such disadvantage from this cause.

Moreover the grid leak itself is effectively across the input circuit, and therefore acts as a shunt resistance replaceable by an equivalent series one (Vol. I).

Damping inevitably reduces the selectivity of the tuned circuit, but it should be noted that the advantages of the grid current detector often compensate for this defect (*cf* N.30).

At any instant the "alternating current resistance" of the grid-filament path is given by the slope of the grid volts/grid current characteristic; under working conditions the effective resistance may be represented by r_g.

Fig. 18 (*a*) shows the simplified circuit, omitting the grid leak and condenser; Fig. 18 (*b*) is the simplified equivalent circuit.

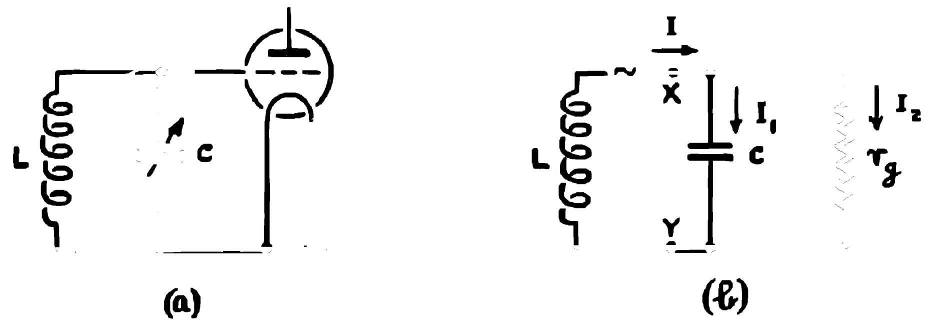


FIG. 18.

It is shown below, that the equivalent series resistance (R_s) introduced into the tuned circuit is approximately given by $R_s = L/Cr_g$. Mathematically, this may be done by finding the effective impedance presented by the circuit to the right of a dividing line through X and Y. This is given by

$$\frac{1}{Z} = \frac{1}{-\frac{j}{\omega C}} + \frac{1}{r_g} = j\omega C + \frac{1}{r_g} = \frac{jr_g\omega C + 1}{r_g}$$

$$\therefore \quad Z = \frac{r_g}{jr_g\omega C + 1} = \frac{r_g(1 - jr_g\omega C)}{1 + r_g^2\omega^2C^2}$$

$$= \frac{r_g}{1 + r_g^2\omega^2C^2} - \frac{jr_g^2\omega C}{1 + r_g^2\omega^2C^2}.$$

The resistive term in this expression gives the **equivalent series resistance,** hence,

$$R_s = \frac{r_g}{1 + r_g^2\omega^2C^2} \doteq \frac{1}{r_g\omega^2C^2} \quad \ldots \text{ a result which is obtained in Vol. I.}$$

Alternatively, using $\omega^2 = \frac{1}{LC}$

$$R_s = \frac{r_g}{1 + r_g^2\frac{C}{L}} \doteqdot \frac{L}{Cr_g}$$

In practice r_g may be as low as 10,000 ohms; including the grid leak (R_1) the total effective parallel load, or "**input resistance,**" may be shown to be approximately half the value of the grid leak, or $R_1/2$ (*cf.* N.28).

An expression indicating approximately the order of magnitude of the **total equivalent series damping resistance** is, therefore

$$R'_s = \frac{2L}{CR_1}.$$

It should be noted that this work does not take account of any damping due to Miller effect (F.11, 32), which may be much bigger than that due to this cause, and may even make the effective resistance negative. (Also *cf.* paragraph 24.)

24. **The Push-Pull Detector.**—In addition to the foregoing single valve detector circuits, it is possible to arrange a rectifying circuit using two valves working in the manner usually known as "push-pull." Using triode valves, the push-pull detector is seen in two basic forms, the anode bend detector, and the grid current detector respectively.

LOWER ANODE BEND DETECTOR.—Fig. 19 (*a*) represents the circuit details of the lower anode bend push-pull detector. The R/F input is applied symmetrically between grid and

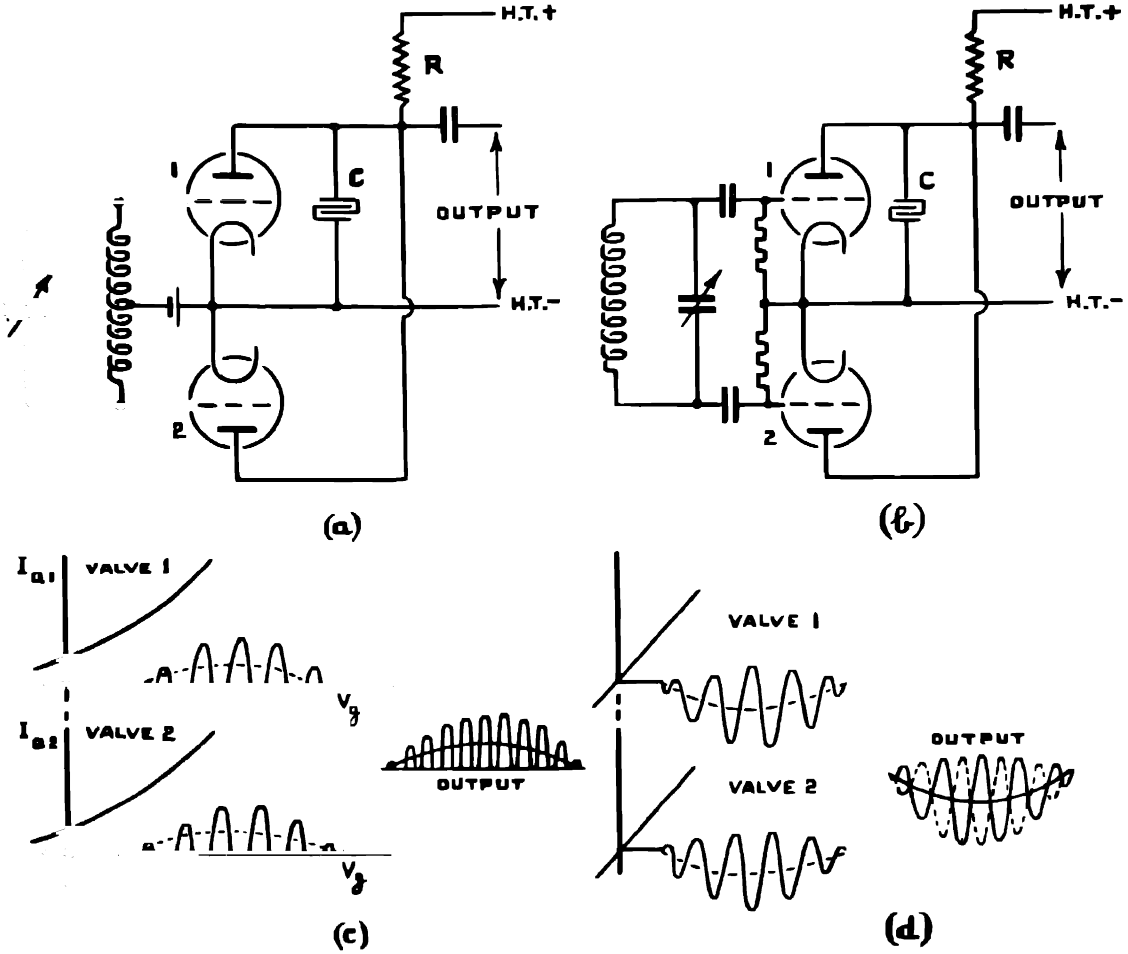


FIG. 19.

filament of [illegible] plied to th[illegible]), the grids being suitably biased negatively in the manner shown. The outputs [illegible] the two valves are connected together, both anodes being fed through the load impedance R.

On receipt of a modulated signal, each valve singly operates as a detector, and Fig. 19 (*c*) represents the R/F pulses of anode current produced by each valve. Anode current only flows during the positive half cycles in each valve, but since both valves are fed through the same impedance, they produce an additive effect in R, giving an output of the form shown in Fig. 19 (*c*). It will be noted that the frequency of the H/F pulses through R appears to have been *doubled*; the circuit, in fact, is one form of "frequency doubling" circuit, a matter which is further referred to in K.46.

The mean anode current through R varies at the modulating frequency, and the latter may be passed on for further amplification if desired. Condenser C provides a by-pass for the R/F component.

A practical example of the use of this circuit appears in Fig. 51 of Section "F."

GRID CURRENT DETECTOR.—Fig. 19 (*b*) represents a push-pull detector circuit using the principle of grid current detection. The output arrangements are similar to those of Fig. 19 (*a*), but the operation of the circuit is somewhat different. As before, a modulated R/F input is applied symmetrically between the grids of valves (1) and (2), but in this case the latter are not adjusted to work at their lower bends. The operation of the circuit may most easily be appreciated by considering its action as a push-pull *amplifier*, disregarding temporarily the rectifying action.

Considering the amplifying action, when the grid of valve (1) is positive, that of valve (2) will be negative; the increase in anode current due to valve (1) is cancelled by the decrease in anode current due to valve (2), and the nett effect in R tends to be nil for the R/F oscillation.

Considering the rectifying action, when either grid is positive, grid current will flow and tend to make the grid go negative. This will cause a decrease in the mean anode current, and since the grid of each valve is positive alternately, the decrease in mean anode current is continuous, and the two effects in R do not cancel out. As the amplitude of the incoming signal increases, the mean anode current is further reduced.

These effects may be represented diagrammatically as shown in Fig. 19 (*d*). The R/F pulses of anode current due to valves (1) and (2) are in anti-phase through R, and tend to cancel each other in the output. If the valves are *perfectly* balanced there will be no R/F current in the common anode impedance. In practical cases the valves are seldom quite the same, and the small R/F content in the anode impedance is by-passed by condenser C, which may be smaller than usual.

The A/F variation of mean anode current, produced by each valve singly, is not cancelled in R, and the effect produced is shown in the output curve of Fig. 19 (*d*).

Theoretically, there is no necessity for grid blocking condensers; the tuned circuit cannot short circuit the grid leak, since it is in series with another equal one (paragraph 6).

Of these two rectifying circuits, the grid current detector is the one offering the more advantages. It gives less damping due to Miller effect (F.11, 32) than the corresponding single valve cumulative grid detector. Miller effect is due to the feed back of H/F energy from the anode circuit to the grid circuit, and produces, in certain cases, an effect equivalent to the introduction of a damping resistance into the input circuit. The latter occurs when the reactance of the output circuit is mainly capacitive, as in the case under discussion. As shown in paragraph 23, the grid current detector suffers considerably from the ill effects due to the damping produced by the flow of grid current; it is therefore important to minimise further damping, and consequent loss of selectivity, due to other causes. In this case Miller effect is minimised, since there is little or no R/F component in the anode circuit.

To offset any advantages we may note the necessity for carefully matched valves, and the greater drain on the H.T. battery than in the case of a single valve detector.

A practical adaptation of the use of this circuit appears in Fig. 12 of Section "T."

25. **Reception of Continuous Waves (C.W.).**—From the foregoing work, and paragraphs 16 and 18 in particular, the asymmetrical current resultant upon the detection of C.W. can be resolved into a high frequency variation about a new steady mean value. Employing power anode (linear)

detection, the process is illustrated in Fig. 20. There is no A/F co[illegible] the rectified current wave form, and the only effect produced in telephones is a click at the beginning and end of a signal.

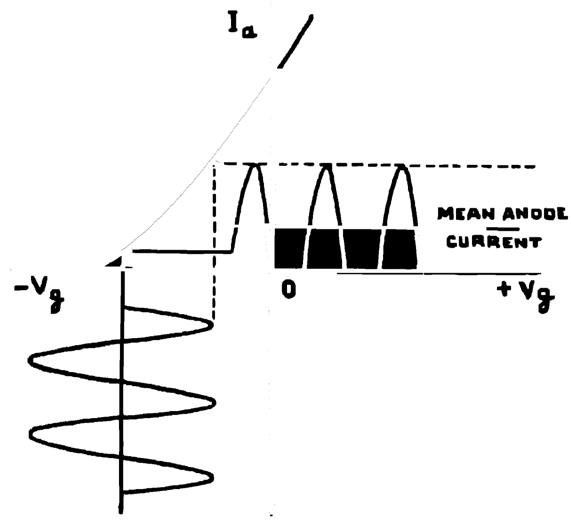


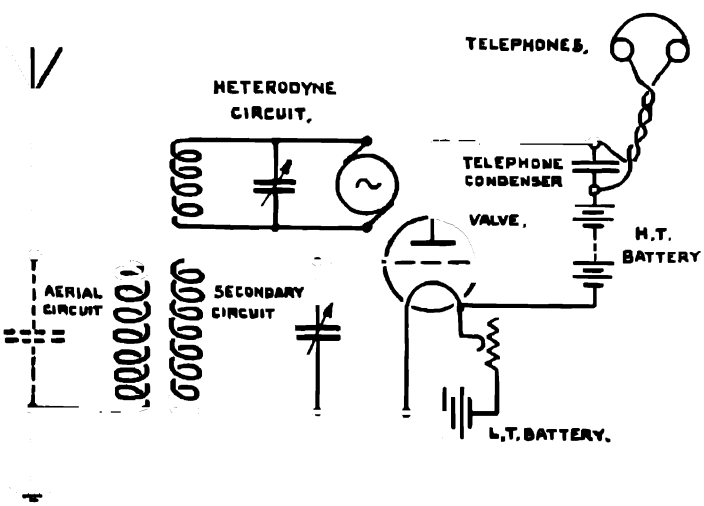
Fig. 20.

The absence of an audible note in telephones is due to the received signal not having any A/F characteristic. In order, therefore, to make the telephone diaphragm vibrate at audio-frequency, it is necessary to break up a long train of continuous waves into groups succeeding each other at audio-frequency, each group giving one pull on the diaphragm as the resultant effect of the asymmetry introduced by the process of detection. This result may be achieved as follows :—

(*a*) At the receiving station, the current in the receiver circuit may be interrupted or varied at audio-frequency by a mechanical method. An arrangement for doing this is called a "tikker," and is generally either an arrangement for breaking and making the circuit, or for periodically mistuning the circuit by breaking or making an additional circuit, or rotating a closed coil coupled to the inductance in the receiving circuit.

(*b*) Heterodyne reception. This method gives more satisfactory results than (*a*), and is described below.

26. **Heterodyne Reception of C.W.**—In this system, a separate local circuit, known as the "heterodyne circuit," is loosely coupled to the detector circuit, as illustrated in Fig. 21. The heterodyne circuit consists of an inductance and capacity, in which a continuous R/F oscillation of suitable amplitude is maintained by some device ; in most cases, the latter is a valve oscillator (Section "K"). For the moment we shall merely assume that this oscillation is set up, and that its frequency is adjustable between wide limits by adjustment of L or C.



SIMPLE VALVE RECEIVER FOR C.W. SIGNALS.

Fig. 21.

When there is no incoming signal, the local oscillator applies an R/F voltage variation of constant amplitude to the detector, but the latter produces no sound in telephones (paragraph 25).

When a C.W. incoming signal is being received, two R/F oscillatory voltages are

simultaneously applied to the detector, one at the frequency of the incoming wave and one at the frequency of the local oscillator.

Now, unless these oscillatory voltages are of exactly the same frequency, they will not rise and fall in time with each other, but will get into step and out of step alternately. The case is not unlike that of two men walking along the street together, one being tall and taking long strides, and the other being short and not taking such long strides as his partner ; if the two keep abreast, they are alternately in step and completely out of step at certain moments ; at intermediate times they are more or less out of step.

The action is illustrated in Fig. 22 ; curve A shows the voltage input due to the incoming wave, curve B showing the input due to, and at the frequency of, the heterodyne circuit. Curve C

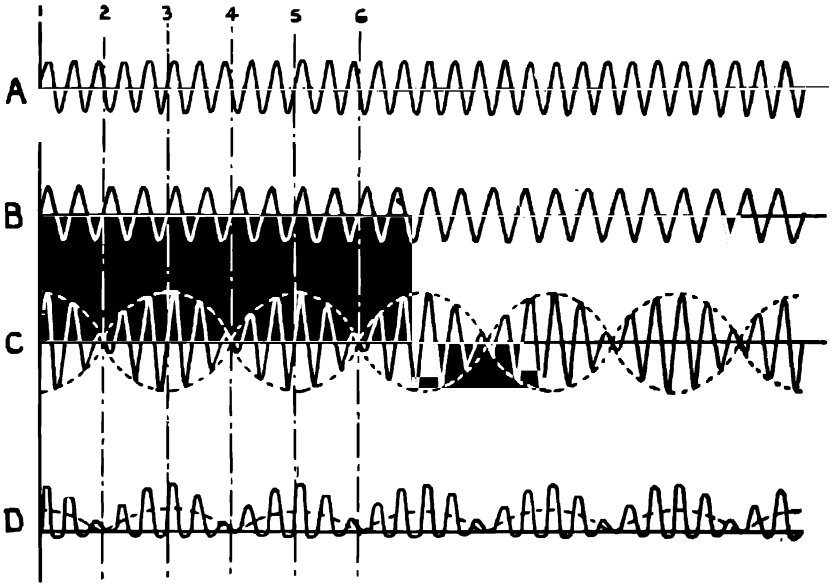


FIG. 22.

shows the combined effect of these two inputs which is applied to the detector. It will be observed that at moments 2, 4 and 6, the two component curves are exactly out of phase ; at moments 3 and 5 they are in-phase, and, therefore, additive.

Curve D shows the rectified current, using lower anode bend detection, and the dotted line through the high frequency variation of current represents the A/F variation of the mean value of the current, on which we may consider the H/F symmetrical oscillations superimposed ; this A/F component actuates the telephones. Each of the sections or groups, into which the wave form of C is divided, gives one variation in the pull on the diaphragm, instead of the constant additional pull which would be the result of detecting either wave form A or B separately.

The wave form illustrated in C is similar to that familiar in acoustics, where a throbbing effect is heard between two organ pipes of nearly the same frequency which are energised simultaneously. The coming in and out of step in such cases is known as the " **beat effect.**"

27. **The Beat Frequency.—The frequency of the beats of voltage** illustrated in curve C, which is also the frequency of the note heard in the telephones, **is given by the difference of the frequencies of the two component R/F waves A and B.** This may be seen from the curves given. Between moments 2 and 4 there are five complete cycles in curve A and four complete cycles in curve B, during which time curve C, the resultant, passes through one complete set or group of variations.

NUMERICAL ILLUSTRATION.—To take figures, the following table gives the beat frequency corresponding to various adjustments of the heterodyne circuit, if a wave at a frequency of 48,000 cycles per second is being received :—

Incoming Wave.	Heterodyne Circuit.	Note heard, or Beat Frequency.
Frequency in cycles/sec.	Frequency in cycles/sec.	Frequency in cycles/sec.
	49,000	1,000
	48,750	750
	48,500	500
	48,250	250
48,000	48,000	0
	47,750	250
	47,500	500
	47,250	750
	47,000	1,000

From an inspection of this table we may deduce that when the frequency of the heterodyne circuit is the same as that of the incoming signal, no sound is heard in the telephones. There will be a space on either side of this point where the beat note is so low as to be inaudible. This occurs if the beat frequency is less than about 20 cycles per second. This space is called the "**Dead Space.**"

As the LC value of the heterodyne circuit is increased or decreased relative to the LC value of the signal, a note of increasing pitch is heard. Eventually the note becomes so high as to be beyond the limit of audibility.

The same note can be obtained at two points, *i.e.*, points at equal differencs of frequency ab and below the dead space, *e.g.*, in the above example a 500-cycle note is heard when the heterodyne circuit is set to 48·5 or 47·5 kc/s.

The great advantage of the heterodyne system of reception can now be seen—that the note is absolutely under the control of the receiving operator. By a slight variation of the heterodyne condenser he can adjust the note in his telephone to any pitch that suits his hearing.

In addition, it is very easy to get over interference by **over-reading.**

For example, if an operator wished to receive a signal of 100 kc/s. frequency, and interference was experienced at 99 kc/s., he could either—

(1) Set his heterodyne to 100·7 kc/s. and read the signal as a 700-cycle note, while the interference would be heard as a 1,700-cycle note (which would be easy to over-read).

(2) Set his heterodyne to the frequency of the interference, or close enough to it to be inside the dead space, in this case to 99 kc/s., so giving the wanted signal as a 1,000-cycle note, while the interference would be inaudible.

★28. **Mathematical Note on the Heterodyne Reception of C.W.**—It is possible to give an accurate proof of the statements based on graphical considerations in paragraph 26.

Suppose that the E.M.F. introduced into the detector circuit by the incoming signal is $a \sin \omega_1 t$, and that due to the local oscillation is $b \sin \omega_2 t$.

The combined E.M.F. is $a \sin \omega_1 t + b \sin \omega_2 t$.

Let $(\omega_1 - \omega_2)$ be written ω.

Then the combined E.M.F. is

$$\begin{aligned} & a \sin (\omega_2 + \omega)\, t + b \sin \omega_2 t \\ &= (a \cos \omega t + b) \sin \omega_2 t + (a \sin \omega t) \cos \omega_2 t \\ &= A \sin (\omega_2 t + \theta), \end{aligned}$$

where

$$A^2 = (a \cos \omega t + b)^2 + (a \sin \omega t)^2 = a^2 + b^2 + 2\,ab \cos \omega$$

and

$$\tan \theta = \frac{a \sin \omega t}{\cos \omega t + b}.$$

The **resultant** can therefore be considered as a sine curve of **varying amplitude A.**

Since $A = \sqrt{a^2 + b^2 + 2ab \cos \omega t}$, it reaches a maximum value every time $\cos \omega t = +1$, *i.e.*, at a frequency given by $\frac{\omega}{2\pi}$.

The separate frequencies of the two waves are given by $\frac{\omega_1}{2\pi}$ and $\frac{\omega_2}{2\pi}$, and their difference is $\frac{\omega_1 - \omega_2}{2\pi} = \frac{\omega}{2\pi}$.

Therefore the frequency with which the amplitude of the wave form in curve C reaches its maximum, *i.e.*, **the beat frequency,** is **the difference of the two separate radio-frequencies.**

The range of variation in the amplitude of the combined wave-form is from $(a + b)$, when $\cos \omega t = +1$, to $(a - b)$, when $\cos \omega t = -1$.

When $a = b$, as in the figure given, the range is from $2a$ to zero.

It may further be observed that, in this case, the value of θ becomes

$$\tan^{-1} \frac{\sin \omega t}{\cos \omega t + 1} = \tan^{-1}\left(\tan \frac{\omega t}{2}\right) = \frac{\omega t}{2},$$

so that the combined E.M.F. may be written

$$A \sin\left(\omega_2 + \frac{\omega}{2}\right) t = A \sin\left(\omega_2 + \frac{\omega_1 - \omega_2}{2}\right) t$$

$$= A \sin \frac{\omega_1 + \omega_2}{2}$$

where $A = \sqrt{2a^2 + 2a^2 \cos \omega t} = 2a \cos \frac{\omega}{2} t$, and therefore

(1) The frequency of the **radio**-frequency oscillations contained in curve C is the mean of the separate radio-frequencies ;

(2) The amplitude variation is simple harmonic.

If a is not equal to b, the radio-frequency in the resultant wave form is itself a varying quantity.

29. **Mathematical Note on Square Law Detection of Heterodyne C.W. Oscillations.**—It is possible to apply the same method of investigation already used in paragraph 18. In this case, the total voltage applied to the grid at any instant can be represented by the expression $V_0 = V_s \sin \omega_s t + V_h \sin \omega_h t$, where V_h and ω_h correspond to the amplitude and frequency of the heterodyne oscillation. On substitution in the formula for I_a, it is found that the mean value of the current is increased by **a steady amount** $\frac{c\,V_s^2}{2} + \frac{c\,V_h^2}{2}$ which does not, of course, produce a note in the telephones. In addition there are **radio-frequency terms,** which are by-passed by the condenser, and **a slow variation** of the mean value, represented by a term of the form

$$c\,V_s\,V_h \cos(\omega_s \sim \omega_h)t.$$

The frequency of this variation of current is therefore the beat frequency, and its amplitude is proportional to the rate of change of slope of the curve, and to the product $V_s\,V_h$. When it is at an audio frequency, this variation affects the telephones. **The audible response is therefore proportional to the product of the amplitude of the input signal and the amplitude of the heterodyne oscillation.** The signal strength increases with increase of the latter up to the point at which the detection process can no longer be regarded as obeying a square law. If the heterodyne amplitude is increased further, it may be considered that linear detection replaces square law detection, the A/F signal strength then depending chiefly upon the first power of the amplitude of the incoming oscillations. There is little advantage in increasing the heterodyne amplitude beyond this point, and, in fact, a point may be reached at which the rectified signal strength again decreases ; the input signals would then be so large that upper anode bend detection partly cancels the effects of lower anode bend rectification.

Since the value of the rectified current is proportional to the product of the amplitude of the incoming signal voltage and that of the local oscillation, it is clear that, for small applied signal voltages, the **sensitivity of the receiver** is much greater for heterodyned C.W. than for signals modulated at the transmitter. This is one of the chief reasons accounting for the popularity of C.W. transmission, and the greater ranges which were obtained by its introduction.

It will be observed that the curves of Fig. 22 treat the case in which the amplitude of the local oscillation is approximately equal to that of the incoming signal. The foregoing work indicates that this is hardly desirable, and, in practice, the heterodyne oscillator supplies an input considerably greater than that of the signal. For this reason, practical results tend to approximate more to linear detection than to square law. In this connection it is interesting to note that it can be shown that square law operation gives distortionless heterodyne detection ; linear law detection produces many harmonics of the beat frequency.

30. **Reception of C.W. at H/F.**—It is important to note that, when working with H/F signals, great constancy of frequency is required at the transmitting end ; a small variation in the LC value of the transmitting circuit, with a fixed adjustment of heterodyne at the receiving end, will cause a big variation in the note heard in telephones. For this reason, heterodyne reception of H/F signals can only be satisfactory if the frequency of the latter is master controlled (*cf.* K.37).

31. **Comparison between C.W. and Spark.**—From the point of view of reception the following contrasts may be drawn :—

(*a*) C.W. reception requires the production of a local oscillation, which is unnecessary with spark or I.C.W. as they are modulated at the transmitter.

(*b*) Spark has the advantage that the transmitting station is defined by its note, or spark train frequency ; but the fact that the operator can adjust the note to suit his ear and to cut out interference, when receiving C.W., more than compensates for the loss of a characteristic note inherent in the signal itself.

(*c*) Circuits can be made much more selective for the reception of C.W., as any wave form which is modulated at the transmitting end may be regarded as being in itself the resultant of several C.W. oscillations at different frequencies. The use of too selective a circuit with spark reception would reduce signal strength by eliminating a lot of the energy in adjacent frequencies. The point will be referred to again under radio-telephony. (Also *cf.* A.1).

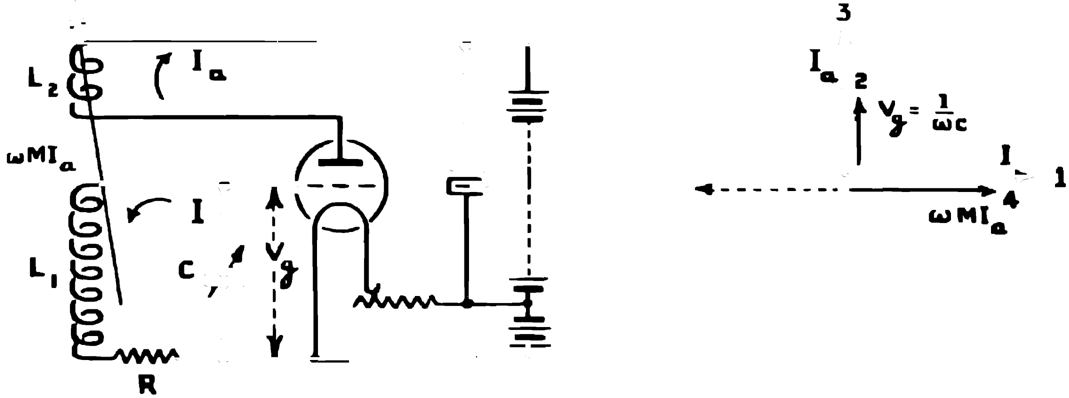
(*d*) Receiving apparatus, equipped with a heterodyne circuit, can be made more sensitive as regards reception of C.W. than when receiving spark.

32. **Design of Heterodyne Oscillators. Further Use of the R/F Component.**—As pointed out in paragraph 3, and subsequently, the anode current undergoes not only the A/F variation which operates the telephones, but also an R/F variation, which is by-passed by the telephone condenser, or otherwise passed to earth.

This R/F component is mainly of the same frequency as that of the incoming wave, but contains also higher harmonics. It **may be usefully employed** by causing it to pass through an inductance in the anode circuit, which is coupled to the inductance L_1 of the input circuit. A radio-frequency oscillatory voltage will then be induced in L_1. It will be shown below, that this induced voltage can be arranged to be in phase with the current flowing in the LC circuit, and hence represent **an introduction of power into the circuit.** If the power introduced is sufficient to balance the damping losses present in the circuit, an oscillatory current started in the latter will not die away, but will be maintained continuously at the same amplitude ; if the power introduced is less than the rate at which energy is being wasted in damping losses, the oscillation will not be continuous, but will be of greater amplitude than if no anode coupling coil were used. In the case of a damped wave train it would endure for longer.

In practice, both of the above conditions may be usefully applied ; the first is used in the design of **oscillators and heterodyne units,** and the second is that used to increase, or prolong, the effects of an incoming signal, and is known as **regenerative amplification, retroaction,** or simply **reaction.** The inductance in the anode circuit is known as the **reaction coil.**

33. **Reaction necessary for Self-Oscillation.**—Fig. 23 (*a*) represents a simple circuit **having a reaction coil** L_2 in the anode circuit, the direction of winding of the latter being the same as that of the grid coil L_1 (*cf.* K.4).



(a) (b)

FIG. 23.

Suppose an oscillation is set up in the LC circuit at its own natural frequency, so that the R.M.S. value of oscillatory current in it is I, in amperes. Energy will be dissipated in the circuit at a rate given by I^2R, in watts, and, with no reaction effect, the oscillation would die away at a rate depending on the damping factor $R/2L_1$ (Vol. I).

In this case, however, the oscillatory current I sets up an oscillatory voltage across the condenser C of value $V_g = 1/\omega C$, in volts, and this voltage leads 90° on the current.

The vector relations are shown in Fig. 23 (*b*), the numbers at the ends of the vectors representing the order in which they may conveniently be studied, or the diagram developed.

Assume the initial oscillatory current of R.M.S. value I begins as soon as the H.T. voltage is switched on when the filament is alight; this is given by vector (1). This current lags ***about*** 90° behind the voltage developed between grid and filament, which is the vector V_g and is the ***back*** *E.M.F.* developed across the condenser C. This grid **input voltage** gives rise to an oscillatory anode current I_a of the same frequency which, in practice, is almost in phase with it; for simplicity the vector (3) is here shown in phase with V_g. The value of the anode current variation is given by $g_m V_g$, in amperes, if one assumes that the reaction coil is small enough for us to neglect its effect on the anode current changes, and therefore to derive the value of the current variations from the **static characteristic** curve.

Owing to the mutual inductive link between the anode and grid circuits, an E.M.F. will be injected into the latter, which will either lead or lag on I_a by exactly 90°, depending upon the sign of the mutual link. Its value is given by ωMI_a in volts, and will be in phase or 180° out of phase with the current I in the input circuit, according to the direction of the windings of the separate coils L_1 and L_2, and the way in which they are joined up (*cf.* Appendix "E").

In this case, vector (4) is drawn for **the self-oscillatory case,** the one in which the **injected volts** are either in phase with the current I that was supposed to start the action, or have a large component in phase with it. When the anode current supplies energy in phase with the current I, a self-supporting oscillatory system is obtained; a simple vector test or criterion for self-oscillatory conditions has thus been developed.

In symbols, when the injected volts and current I are in phase, this represents an introduction of energy into the circuit at the rate of $(\omega MI_a)I$, in watts. But, on account of damping losses, ***energy is being dissipated*** in the circuit at the rate of I^2R, in watts. Therefore the nett loss of energy is at the rate of $(I^2R - I \times \omega MI_a)$ watts.

Now $I_a = \frac{g_m}{\omega C} I$, and so $\omega M I_a \times I = \frac{g_m M I^2}{C}$. Hence the nett rate of loss of energy is $I^2 \left(R - \frac{M g_m}{C}\right)$ watts. If this is a positive quantity, *i.e.*, if $R > \frac{M g_m}{C}$, oscillations set up in the LC circuit will die away, but at a slower rate, because the effective resistance of the circuit may be considered to be reduced from the value R to the value $\left(R - \frac{M g_m}{C}\right)$. The energy derived from the reaction coil has partially neutralised the damping of the oscillatory circuit. If, however, $R = \frac{M g_m}{C}$, the effective resistance of the circuit and the damping factor both become zero, and oscillations set up in the circuit are maintained at a constant amplitude.

If R is less than $\frac{M g_m}{C}$, **the effective resistance is negative,** and oscillations will increase in amplitude up to a point where limiting conditions, hitherto neglected, come into play, and prevent the increase going on indefinitely. These will be examined in the next paragraph.

In the latter two cases we have utilised the valve as a medium for maintaining self-oscillations in an oscillatory circuit ; in other words, the valve has been acting as **a generator of C.W. oscillations.** The actual circuit shown is not suitable for high power transmitters, but it is very useful for producing the small local oscillations necessary to beat with incoming signal C.W. oscillations. and to act as a heterodyne unit.

The conditions for maintenance or otherwise of oscillations can be stated in a slightly different form. They depend on the relative values of R and $\frac{M g_m}{C}$. Since it is the closeness of coupling and hence the mutual inductance, M, between L_2 and L_1 which is the adjustable variable, it is usual to express the conditions in the form of a comparison between M and the other quantities involved.

This is easily seen to be as follows :—

If M is greater than $\frac{CR}{g_m}$, oscillations increase in amplitude.

If $M = \frac{CR}{g_m}$, oscillations are maintained at constant amplitude.

If M is less than $\frac{CR}{g_m}$, oscillations die away, but at a less rapid rate than if no reaction were introduced.

34. **Conditions Limiting the Amplitude of Self-Oscillations.**—When $M > \frac{CR}{g_m}$, oscillations increase in amplitude, because the effective resistance of the circuit is negative, *i.e.*, the damping factor, $\frac{1}{2L_1}\left(R - \frac{M g_m}{C}\right)$, is negative. The following factors limit the growth in amplitude to a definite amount.

(i) So far we have assumed the slope "g_m" to be that of the static characteristic. If, however, the grid voltage variations increase to large values and give correspondingly large changes in anode current, the effect of the reaction coil must be taken into account. The result is that the curve which represents the actual anode current plotted against grid voltage is **the dynamic characteristic,** instead of the static characteristic. Now the average slope of the dynamic characteristic is always less than that of the static, and hence "g_m" has a lower value than that hitherto assumed and the preponderance of M over $\frac{CR}{g_m}$ is not so large. This, however, does not account for the ultimate cessation of the increase in amplitude, but it describes more correctly the *inequality* hitherto found.

(ii) Again, we have not taken into account the fact that large grid voltage variations are bound to give a considerable flow of grid current during the time the grid is positive, and so to

produce an increased damping effect on the oscillatory circuit. In other words, R increases above the value hitherto assumed, which was simply taken as the ohmic resistance in the LC circuit.

(iii) Finally, there is the most important condition from the point of view of limitation, that the anode current curve, whether static or dynamic, is not continuously straight, but reaches points, both at the top and bottom of its straight part, where it flattens out and becomes horizontal. Therefore, after the oscillatory grid voltage reaches a certain amplitude, it is incorrect to say that the corresponding amplitude of oscillatory anode current is given by the product of the mutual conductance, g_m, and the amplitude of grid voltage swing. As soon as the increasing V_g variations reach either bend, further increase of I_a in that direction is prevented.

This is shown in Fig. 24, where increasing V_g variations are accompanied by I_a variations of proportionate amount, until the latter are limited by saturation and zero values We might express

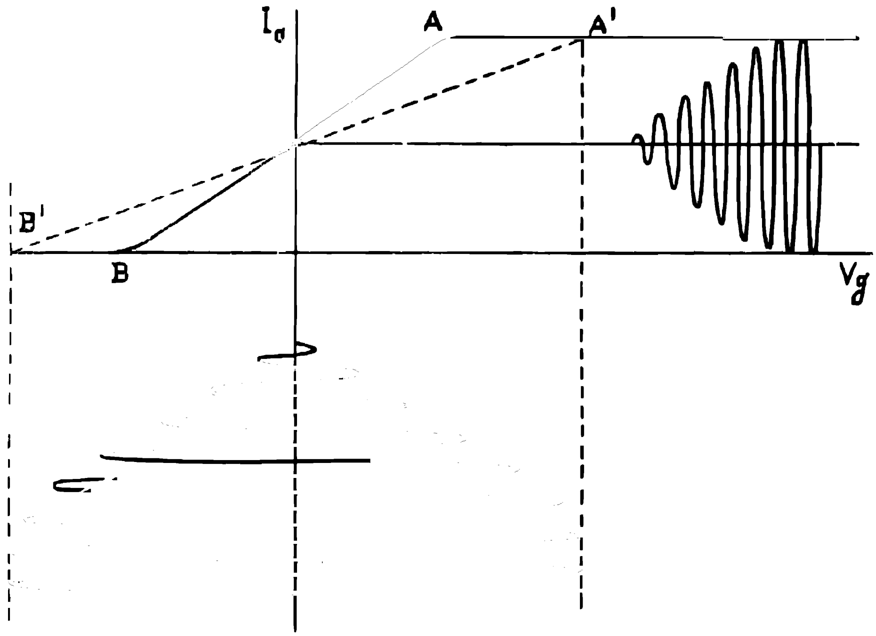


FIG. 24.

this fact by saying that, after the amplitude of the grid swing goes beyond the straight portion of the characteristic, the ratio of anode current variation to grid voltage variation is given by the slope of the line A′ B′, which is less than g_m, the slope of AB. In other words, the slope of A′ B′ (*cf.*F.4) should be substituted for g_m in the expression $\frac{CR}{g_m}$, and, as the amplitude of oscillation increases the denominator of this expression will get steadily less, and the value of the expression greater. Hence, although at the beginning of operations the inequality $M > \frac{CR}{g_m}$ is true, an amplitude of oscillation is reached sooner or later at which it is no longer an inequality, but an equality, and then oscillations will be continuously maintained at this amplitude. The same result may be obtained by comparing the damping losses with the power reintroduced into the circuit. The damping losses are given by I^2R, and the power introduced by reaction is the product of I and the voltage induced in the oscillatory circuit. This voltage is given by ωM multiplied by the variation in anode current, which cannot exceed a certain value, and hence, as I increases, the power introduced into the circuit is ultimately proportional to I. The damping losses are, however, proportional to I^2, and therefore increase at a greater rate, until a point is reached at which there is a state of equality.

Hence, if the coupling is greater than the critical value necessary for self-oscillation, oscillations

will increase in amplitude until a condition is reached when limitation of the value of oscillatory anode current flowing has the effect of making the coupling just sufficient to maintain oscillations at this definite amplitude.

35. **Reaction Receivers.**—The circuit of Fig. 23, with the addition of telephones across a condenser in the anode circuit, and a coupling to a receiving aerial, together with arrangements for securing anode or cumulative grid rectification, could be used for the reception of modulated incoming signals. For this purpose, the coupling of the reaction coil is adjusted to be less than that necessary to set up oscillations in the LC circuit, so that its effect is to diminish the effective resistance present in the circuit, and to give larger amplitudes of oscillatory current, and of voltage applied to the grid for a given signal voltage induced into the circuit from the receiving aerial. The effect is shown in Fig. 25, taking, for simplicity, the case of a C.W. oscillation. The full curve represents the amplitude of the signal without employing regenerative amplification.

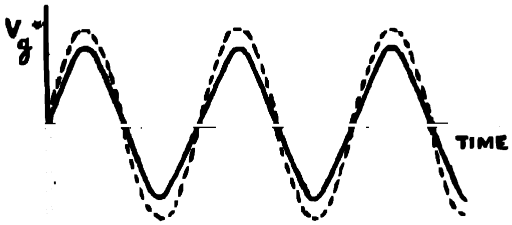


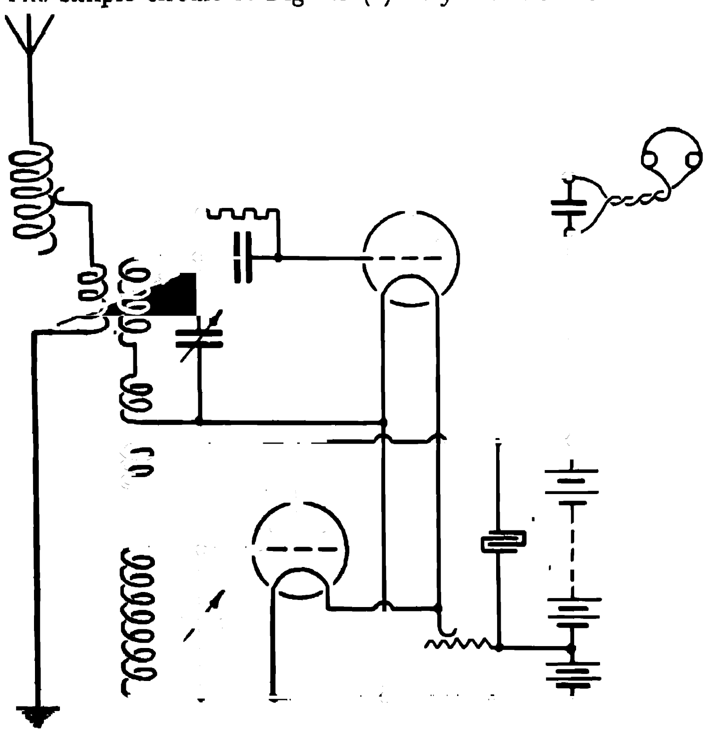
FIG. 25.

If the coupling is increased too much, a continuous oscillation will be set up, quite irrespective of the incoming wave; this spoils the note of an I.C.W. signal, and distorts speech in the case of R/T.

It should also be noted that the **use of reaction,** by decreasing the effective damping, **increases the selectivity of the receiver** ; in W/T this may be good, but for R/T it may produce distortion (N.1, 31).

Reaction receivers are referred to in greater detail in F.44 and subsequent paragraphs.

36. **Complete Circuit of a Receiver Employing a Separate Heterodyne Oscillator.**—The simple circuit of Fig. 23 (*a*) may also be used as a means of generating the local oscillations necessary for heterodyning. In this case the coupling must be sufficient **for self-oscillations** to be set up. The circuit in which the heterodyne oscillation is generated must, of course, be mistuned from the frequency of the incoming signal to the detector valve, so as to give the necessary beat note. A complete circuit for the reception of C.W., using a detector valve and a separate heterodyne unit, might be arranged as in Fig. 26. The anode coil of the heterodyne oscillator is mutually coupled to the input inductance of the detector.



HETERODYNE AND DETECTOR.

FIG. 26.

37. **Autodyne or Autoheterodyne Circuit.**—The same complete circuit as that used for regenerative amplification, and re-drawn below, may be used conveniently both to heterodyne an incoming continuous wave, and also to rectify the resultant "beats" due to the interaction

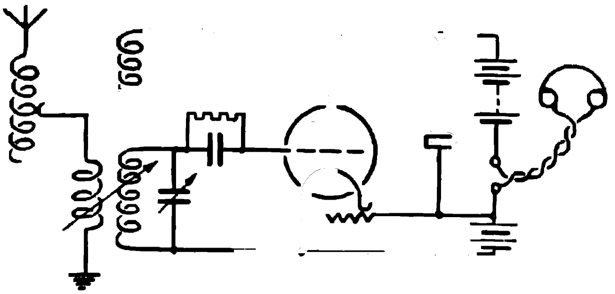


FIG. 27.

between the incoming wave and the heterodyne oscillation.

The necessary conditions in this case are that the **coupling must be sufficient to give self-oscillation, and the LC circuit between grid and filament must be mistuned** from the incoming oscillations. The circuit is known as an **autodyne** circuit, and performs the functions of local oscillator and detector.

Action.—Before a signal arrives, the valve is maintaining oscillations at a frequency f_1 determined by the LC value of the oscillatory circuit, and of amplitude limited by the conditions of paragraph 34. The presence of the grid leak and condenser, necessary for cumulative grid detection, simply means that the grid's steady potential, about which oscillations take place, is somewhat negative (*See* paragraph 19). A steady current is passing through the telephones, the radio frequency variations of anode current being by-passed by the condenser.

An incoming C.W. oscillation will set up induced voltages and currents in the LC circuit, at a frequency f_2, which differs from f_1 by the beat note. It should be noted that the amplitude of oscillatory current and resulting voltage between grid and filament at frequency f_2, is less than if the LC circuit were tuned to resonance with f_2. We have now two high-frequency oscillatory voltages applied to the grid at frequencies f_1 and f_2, and, being superimposed, their resultant effect is a high-frequency oscillation whose amplitude varies at the " beat " frequency, $f_1 — f_2$.

This will result in the mean potential of the grid varying at the beat frequency, as in the ordinary theory of cumulative grid detection, and hence an audio frequency rise and fall in the mean anode, battery and telephone current, giving an audible note in the telephones.

The best condition of combined heterodyning and detecting, is when the self-oscillation is not unduly strong compared with the incoming signal, and this condition may be obtained by reducing the reaction, the anode voltage, or the filament current, until a small amplitude of self-oscillation is just maintained. Both filament current and anode voltage control operate by diminishing slightly the slope of the mutual characteristic (B.26).

38. **Comparison of Autodyne and Separate Heterodyne.**—The **advantages** of the autodyne arrangement over the separate heterodyne can be summed up in the statement that it involves one circuit instead of two, occupies less space, and needs less adjustment. The **disadvantages** are, however, considerable.

(*a*) It is difficult to control effectively the strength of the local oscillation. If local oscillations are to be maintained, these will increase until the variations of grid voltage cover a large portion, if not all, of the straight part of the (I_a, V_g) characteristic. With a separate heterodyne, the amplitude of the local oscillations introduced into the detector circuit can be adjusted by varying the distance apart of the coupling coil of the heterodyne circuit and the inductance in the detector circuit.

(*b*) In order to obtain the " beat " note, the LC circuit in the autodyne receiver must be thrown out of resonance with the incoming wave, whose effect is consequently weakened. This loss is particularly serious in the case of lower frequencies, as a given frequency difference represents a greater percentage of mistuning. For these it is generally best to use a separate heterodyne.

(*c*) With an autodyne, it is difficult to prevent radiation of the locally produced oscillation from the aerial, which must, from the nature of the circuit, be coupled to the inductance of the oscillatory circuit ; interference is consequently caused to other receiving sets in the neighbourhood working on the same or adjacent frequencies. This difficulty may be overcome to some extent in the case of a multi-stage circuit (*i.e.*, amplifiers + detector), by setting into oscillation a tuned circuit in the receiver at a later stage, where it is not in close contact with the aerial.

Similar trouble arises when a receiving set, not intended for C.W. reception, generates self-oscillations through reaction being pushed too far, or from various factors which are liable to give the same effect, and which will be examined in the section on amplifiers.

39. **Tuning Arrangements of Receiving Circuits.**—With reference to paragraph 2, the foregoing work has been mainly directed towards the study of the detector, and the need for detection of E.M. waves. We can now return to consider, in greater detail, the arrangement of the tuned circuits necessary in order to apply the requisite voltages to the detector.

As previously explained, the main objects of the receiver are to produce sufficient **selectivity** in combination with adequate **audibility.** In some respects these two requirements are inter-dependent.

In regard to audibility, with the help of amplifiers (Section "F") it is possible to bring up any signal, however weak, to audible strength; it is not so important to have an arrangement of receiving circuits which will apply the *maximum* possible voltage to the detector, as to have an arrangement which will give satisfactory selectivity combined with an *adequate* oscillatory input to suit the detector in use.

Selectivity depends on the shape of the response curve of each of the tuned "stages," and also on the number of stages. High selectivity means the avoidance of interference from atmospherics, or from signals at a frequency different from that which it is desired to receive. The overall selectivity of a receiver is a result of the joint action of its various circuits (*cf.* N.64); although a receiver may be efficient as regards amplification, it is useless if it is unselective, because the unwanted interfering signals are amplified together with the wanted one.

The position of the tuned circuits, on the design of which selectivity ultimately depends, is not a matter governed by hard and fast rules; quite often it is a matter of policy based on a number of considerations. In some cases it will be found that the *aerial circuit* is tuned to resonance (*cf.* Fig. 6 and Section "F," Fig. 12), in other cases, perhaps more often, it will be found in the untuned or aperiodic condition; in the latter case, all signals are received indiscriminately, being subsequently passed through a buffer or isolating stage, often a screen grid stage, the requisite signals then being selected by suitably tuned circuits after the buffer stage. Figs. 29 and 35 in Section "F" are particular cases illustrating this point.

In employing most of the older service receivers the aerial was tuned; in many respects it appears obviously better to tune the aerial, so that maximum current may be obtained from the small input voltages applied by the signal. In these early receivers the **tuner** was usually a separate unit (*cf.* Section "F," Fig. 12).

Untuned aerials offer a number of advantages, among which are the following:—

(*a*) In large W/T installations many receivers may be operated from one aerial, provided that suitable means are employed to eliminate interaction between receiver and aerial, and between receivers. This matter is sometimes approached by feeding different receivers through filters from the same aerial (R.27).

(*b*) In H/F work it is found particularly advantageous to use untuned aerials (*cf.* F.43).

(*c*) In the design of commercial broadcast receivers it is usually found better not to tune the aerial in order to render the calibration of the receiver as independent as possible of the circuit constants of the aerial; with the ordinary domestic broadcast receiver, the latter are inclined to vary within wide limits.

40. **Selectivity and Response Curves; "Q."**—In Vol. I it is shown, for the case of a series LC circuit, that **the ratio of the resonant current to the current at a given percentage departure from resonance depends on the value of** $\frac{1}{R}\sqrt{\frac{L}{C}}$. The greater this value becomes, the sharper is the peak of the resonance or response curve. We have, therefore,

$$\text{Selectivity} \propto \frac{1}{R}\sqrt{\frac{L}{C}} = \frac{\omega L}{R} \ldots\ldots\ldots \left(\text{since } \omega^2 = \frac{1}{LC}\right)$$

The fraction $\omega L/R$ has become a popular **criterion of selectivity**; it will be noted that it is the reciprocal of the "**power factor,**" $R/\omega L$ of the coil (*cf.* power factor of condenser).

Power factors employed in coils designed for radio work vary within wide limits; typical values for those used in the highly selective wavemeter circuits vary from about 0·0015 for a small inductance, to 0·006 for one of the order of 1-henry (W.9). In other cases, however, using very sensitive resonance indicators, adequate frequency discrimination in wavemeter circuits has been obtained employing power factors as high as 0·01 and 0·03.

There is also an older criterion of selectivity based upon the **log decrement** $\pi R/\omega L$; in terms of this we have—

$$\text{Selectivity} \propto \frac{\pi}{\text{log. dec.}}.$$

In the case of the R/F circuits of the tuner in Fig. 29 of Section " F," the log decrement of the coils is about 0·038. The value of δ for the note selector circuits used in that receiver varies from 0·05 in the most selective condition, to 0·3 when heavily damped.

The newest and most convenient method, however, employs the fraction of $\omega L/R$, to which the symbol " Q " is usually given; hence, we have

$$Q = \frac{\text{coil reactance}}{\text{coil H/F resistance}} = \frac{\omega L}{R}.$$

In any series LCR circuit, with an applied voltage E, the current I at resonance is given by E/R, and we have

$$\text{Voltage across L at resonance} \doteqdot \omega LI = \frac{\omega LE}{R} = EQ$$

(neglecting the resistance of the inductance L).

The resonant rise in voltage across the inductance, or across the condenser, is thus approximately Q times the applied voltage. For this reason Q is sometimes called the **coil amplification factor**; the bigger Q, the bigger will be the voltage amplification due to resonance, and the **stiffer** will be the tuning of the circuit.

In the Service, the phrase " **stiffly tuned** " implies high selectivity.

Q (or its reciprocal, the power factor) is a particularly useful quantity, since its value tends to remain constant for the same coil over a wide range of frequencies; the latter is due to the fact that the effective radio frequency resistance of a coil is almost proportional to the frequency.

Typical values of Q may be obtained from the typical values of power factors, already quoted; in that case it will be seen that Q varies from 33·3 to 666. In receiving circuits more common values range from about 100 to (say) 200.

In the case of the R/F circuits of the tuner in Fig. 29 of Section " F," the value of δ already quoted gives the figure 83 for Q. Similarly, in the case of the note selector circuits, Q varies from 63 to 10·4.

The above considerations emphasise the importance of this factor in determining the shape of the resonance curves of the circuits composing a receiver; it should be noted, however, that some workers prefer to direct attention to the L/R ratio. Whatever criterion is adopted, it is clear that the selectivity is inversely proportional to the H/F resistance R.

Constant selectivity can only be obtained if Q is constant for all of the circuits; this may not be desirable in all cases, and for the purposes of R/T it is shown that permeability tuning (F.20) has the advantage of providing selectivity which increases with the frequency, the ratio L/R being constant

41. **Tuning the Aerial Circuit.**—As indicated above, the selectivity of a receiver is that due to its various tuned stages, one of which may be the **aerial.** The subject is a complicated one, and further reference is made to it in Section " R."

Considered simply, if the frequency to be received is less than the resonant frequency of the aerial, a **loading inductance,** often called the aerial tuning inductance (A.T.I.), is added in series, in order to increase the LC value and make the fundamental frequency of the *aerial circuit* lower. Conversely, if the frequency to be received is greater than the resonant frequency of the aerial, a **shortening capacity,** often called the " aerial tuning condenser " (A.T.C.), is added in series with

the aerial capacity (σ), in order to decrease the LC value by decreasing the total capacity. Combinations of these methods may be used, as shown in Fig. 28 (*b*); with such an arrangement a large range of wave frequencies can be covered.

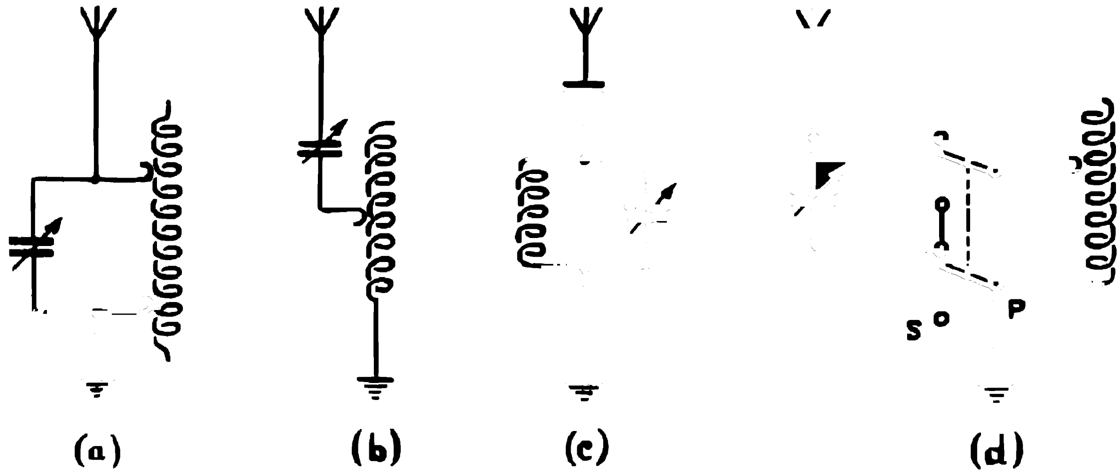


FIG. 28.

For frequencies lower than the resonant one, it is sometimes inconvenient to use more than a certain amount of inductance in series; in this case, the condenser may be joined in parallel with the inductance, so that it is, in effect, in parallel with the aerial capacity, as shown in Fig. 28 (*a*). The total capacity is then the sum of the separate capacities. A further method is shown in Fig. 28 (*c*), using both a series and a parallel condenser, as well as an inductance. The latter circuit has the advantage that the tuning capacity may be made a greater proportion of the whole, thereby giving a greater tuning range. In the case of Fig. 28 (*a*), if the circuit is tuned with a condenser of value 0·0005 μF, it is possible that there may be a permanent aerial capacity (σ) of similar value across the circuit in parallel; in that case the capacity range can only be 2 : 1, giving a corresponding frequency variation of only 0·7 : 1, instead of a more usual 1 : 3 ratio. By means of a double-pole 2-way switch, arrangements (*a*) and (*b*) can be produced from the same inductance and condenser, as shown in Fig. 28 (*d*). Practical cases of the latter **series-parallel condenser** are seen in Section "F" Fig. 12, Section "K" Fig. 21, and elsewhere.

An alternative way of regarding an aerial, is to consider it as a series LCR circuit, being resonant, like an acceptor circuit, when the inductive reactance (ωL) is equal to the capacitive reactance ($1/\omega C$). **The operation of tuning,** consists in adjusting these two reactances to equality. For frequencies above the resonant one, the inductive reactance exceeds the capacitive reactance, and the balance can only be achieved by increasing the latter, an operation which involves making the total capacity smaller. For frequencies below the resonant one, the requisite equality can only be achieved by increasing the inductive reactance, by increasing the inductance. (*Cf.* R.19.)

It is useful here to realise that aerial lengths are more conveniently measured in terms of the wavelength (λ) in use, the latter being obtainable from $c = f\lambda$ (*cf.* R.15).

In Section "R" it is shown that, when working at L/F or M/F, the shape and length of the current *standing wave* (R.14) on an aerial adjusted to resonance, corresponds to a quarter wave form as conventionally drawn. For this reason these simple aerials are often called "**quarter wave aerials,**" and the operation of tuning may thus be regarded in yet another light, as an adjustment of the length of the aerial to make it electrically equivalent to a $\lambda/4$ aerial.

When simple vertical aerials are used at H/F, the length of the aerial may be many times the wavelength in use; many complete standing waves of current may be on the aerial but, in Service cases, the operation of tuning may be regarded as an adjustment of the total length to make the whole circuit *electrically equivalent* to an odd number of quarter wavelengths, *i.e.*, electrically similar to the simple quarter wavelength aerial (*cf.* R.20).

The voltage induced into a receiving aerial by a signal is proportional to the product of the field strength, expressed in microvolts per metre, and the effective height of the aerial (*cf.* R.29); for high signal strength, simple aerials should be made comparable with a quarter wavelength in height. At L/F and M/F, in most practical cases a receiving aerial is only a fraction of a wavelength long.

The **selectivity of a tuned aerial circuit,** considered by itself, is usually low. This is mainly because an aerial has a relatively large total resistance, some of which is *radiation resistance* (R.24); this damping resistance decreases the effective Q of the circuit and broadens the tuning. The response curve is relatively flat in form. This is one of the reasons why it is often better to use the aerial only as a source of energy from which signals may be extracted and applied to circuits in which aerial resistance is not present, and where higher values of Q can be obtained.

When a receiver is attached to an aerial it applies a load to the latter, in many respects similar to the action of an external resistance joined across the terminals of a battery. In that simple case it may be shown that maximum power can be applied to the load when the impedance presented by the load is adjusted to be equal to the internal resistance of the unit delivering power, the aerial in this case (*cf.* R.35). The parallel is a complete one, and many cases of this impedance matching will be found (*cf.* Section " F," Fig. 38).

Simple practical cases of tuned aerials are seen in Section " F," Fig. 12, in the *stand-by position*, and also in the stand-by position of Fig. 29 of the same section, when the isolating stage is not being used.

42. **Selectivity and Coupled Circuits; Mutual Coupling.**—As the foregoing work has shown, the damping of a tuned aerial circuit is always high and its selectivity cannot be very good. If it is desired to make a directly coupled aerial as selective as possible, the coupling to the receiver should be adjusted so that the latter presents a load to the aerial less than the optimum one, *i.e.*, that giving maximum power output (paragraph 41). In general, the requisite selectivity is obtained by passing on the signal to circuits having higher values of " Q," *i.e.*, circuits which are less heavily damped. The directly coupled aerial is only used when searching for signals in what is called the " **stand-by** " position; once a signal has been picked up, the circuit is switched into a more selective arrangement, known as the " **tune** " position. The latter is often provided by a tuned secondary circuit coupled mutually to the aerial circuit, a simple example having already been seen in Fig. 6.

Instead of the receiver being joined across a portion of the aerial circuit, as in Fig. 29 (*a*), it may be joined across a separate circuit, which may consist of an inductance only, or, more generally, an inductance and capacity, the LC value of which is made to correspond with that of the incoming wave; a simple coupled circuit is shown in Fig. 29 (*b*).

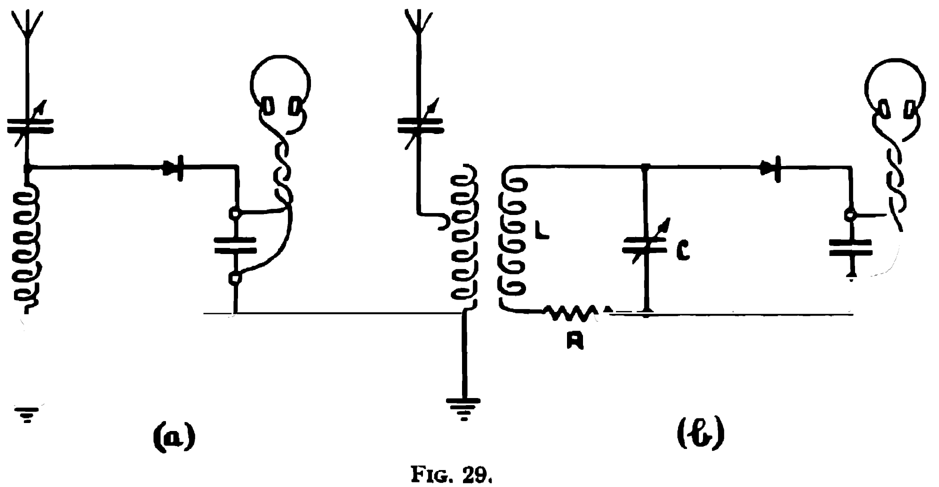


FIG. 29.

The stand-by and tune positions are clearly shown in Figs. 12 and 29 of Section " F."

Oscillatory currents in the aerial set up H/F alternating voltages in the secondary circuit by means of the mutual coupling. It should be understood that **the secondary circuit behaves as a series circuit with respect to such injected voltages.** The voltage introduced into the circuit can be regarded as that of a small alternator joined in *series* with the secondary circuit; if the secondary is tuned to resonance there will be an alternating current in it, of value given by the R.M.S. voltage divided by the ohmic resistance. In the case of Fig. 29 (*b*), the actual voltage applied to the detector device is the voltage across the secondary condenser consequent upon this current flowing.

If the secondary induced voltage is E, the current is $\frac{E}{R}$, and the voltage applied to the detector (V_d) is given by—

$$V_d = \frac{E}{R} \times \frac{1}{\omega C} = \frac{E\omega L}{R} = EQ.$$

The voltage applied to the detector, therefore, may be much greater than E, its actual value depending upon the Q of the circuit.

Because of the fact that there are two tuned circuits instead of one, the coupled secondary circuit arrangement is the more selective. Forced oscillations, set up in the first circuit, the aerial, by incoming waves of non-resonant frequency, produce smaller currents than a resonant wave of the same amplitude (field strength), and since the secondary circuit is also tuned to the resonant frequency, the voltages they induce in it give currents whose value is still further cut down. If the two circuits are equally selective, an unwanted signal the response to which is one-tenth of that of the wanted frequency, will be " ten times down " in the aerial circuit and " one hundred times down " in the secondary one, since the input to the latter is only one-tenth of that of the wanted signal.

If increased selectivity is required, the number of circuits may be increased, as in Fig. 30; usually, one tuned circuit in addition to the aerial circuit is found to be sufficient, and every additional circuit means more complicated tuning and greater losses.

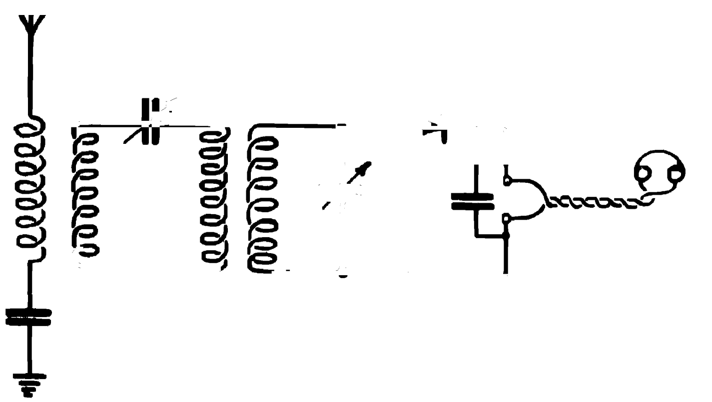


FIG. 30.

Any extra tuned circuits which are introduced between the aerial and the detecting device, or between the aerial and the R/F amplifier, are often regarded in Service sets as a separate unit of the complete receiver (*cf.* Section " F," Fig. 29), and is usually called the " **tuner.**" Coupled circuits are often found having the aerial in the untuned or aperiodic condition.

Fig. 31 (*a*) represents a mutually coupled system, and Fig. 31 (*b*) represents its counterpart employing an auto-transformer coupling.

Considered electrically, each of these coupled circuits may be regarded as a form of R/F transformer coupling. Certain particular circuits are treated in F.19, the case of F.19 (*d*) being approximately that of Fig. 31. With suitable design, transformer voltage step-up effects may be produced, and the effect of the aerial capacity in controlling the frequency of the circuit, may be appreciably reduced.

As with all transformers, any load applied to the secondary produces a corresponding **reflected load** in the primary, in this case the aerial circuit. The magnitude of the reflected load is controlled by the percentage coupling and the transformer ratio.

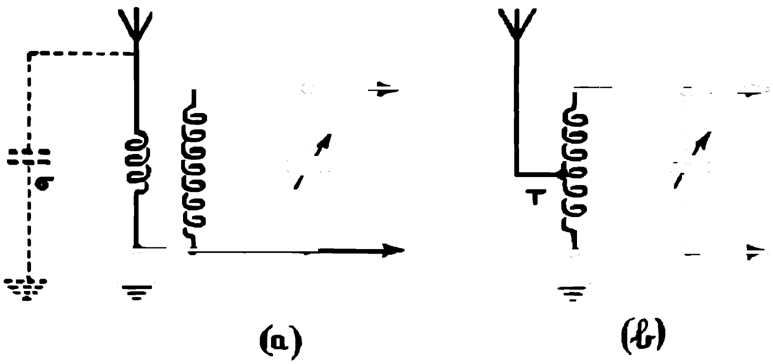


FIG. 31.

In view of these considerations, the position of **the aerial tapping** in Fig. 31 (*b*) is a matter of importance. If T is at the top, the circuit is equivalent to Fig. 28 (*a*) ; if T is at the bottom of the coil, no signals at all will be received. Maximum coupling is obtained with the tap T at the top of the coil, and the lower the tap the weaker the signals but the higher the selectivity, since the reflected load becomes progressively less. For high selectivity the tap should be low down, but for high signal strength it should be higher ; in practice, its position is, therefore, a compromise, and it is usually fixed.

It should be noted that Fig. 29 (*b*) might also be considered to represent an aperiodic aerial coupled to a secondary circuit. In that case the aerial condenser is not used to tune the aerial ; it may be considered as a small condenser, the object of which is to reduce the effective capacity of the aerial, and to increase the selectivity of the aerial circuit by keeping the load on it small (*cf.* N.63). Such an aerial circuit will have nett capacitive reactance, any decrease in which will produce a corresponding increase in the strength of the signals ; the aerial load will increase, and the circuit will become less selective. The aerial is most selective when the aerial condenser is small and, logically, would be at its best when the capacity of the condenser is zero. That would correspond to using a receiver without an aerial, an arrangement which might result in zero signal strength.

43. **Percentage Coupling; Selectivity and Strength of Signals.**—As shown above, selectivity depends not only upon the value of Q for each of the tuned circuits, and on their number, but also on the percentage coupling which exists between them, the aerial and the secondary circuit in these simple cases.

STRENGTH OF SIGNALS.—Whatever the nature of the coupling, if it is *loose* the aerial is unable to hand over to the secondary circuit much of the energy which it absorbs from the passing E.M. waves, and most of it is dissipated in I^2R losses in the aerial, while some is even re-radiated into space.

As the coupling is increased, the secondary circuit takes a greater percentage of energy from the primary, its effect being equivalent to a reflected resistance the value of which can be estimated by ordinary transformer theory.

If the coupling is increased beyond a certain optimum point, the increased resistance in the aerial circuit, due to the reflected load, cuts down the aerial current to such an extent that the power in the secondary will be less than its value for a looser coupling. The case of the *over-coupled* untuned aerial may be investigated by a method similar to that of F.19 (*d*) or paragraph 336, Vol. I, and it may be shown that the coupling for maximum secondary current is obtained when $\omega^2M^2 = Z_1{}^2R_2/R_1$, where M is the mutual inductance, Z_1 the aerial impedance and R_2 the secondary resistance ; a simultaneous condition for optimum coupling is that $\omega^2M^2 = X_2Z_1{}^2/X_1$.

If the primary is tuned, a double frequency effect is set up with tight coupling, and, by the theory of forced oscillations and coupled circuits, the circuits will be resonant to two frequencies,

one higher and one lower than the actual frequencies to which they are separately tuned. Under these conditions, the aerial and secondary circuits form two tuned coupled circuits. If an E.M.F. is impressed on the primary, of frequency equal to that to which the two circuits are tuned, the current in the secondary will have a maximum value for a certain critical or **optimum coupling,** and for tighter or looser coupling than this the current will be reduced. In F.19 (*e*) it is shown that the optimum coupling is given by the expression $K = 1/\sqrt{Q_1 Q_2}$. These optimum conditions occur when the resistance reflected into the primary from the secondary is equal to the initial primary resistance (paragraph 41).

SELECTIVITY.—As observed in paragraph 42, audibility and selectivity are interdependent. A decrease of coupling below the optimum value increases the selectivity but decreases the strength of signals.

For W/T purposes, it is generally advisable, therefore, to use a coupling which is less than the critical value, and so to gain selectivity, since the strength of signals can always be increased by amplification. These effects are illustrated in Fig. 32.

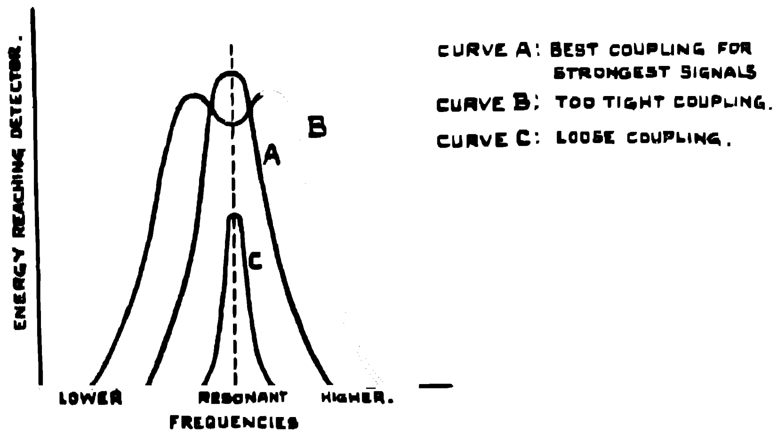


FIG. 32.

Curve A illustrates what happens with critical coupling. The selectivity obtainable is illustrated by the slope of the curve.

Curve B shows what happens with coupling so tight that, with two tuned circuits, a double frequency effect is set up. The response at resonance is less than for curve A, and peak values are obtained for frequencies a little higher and a little lower than the resonant one. The curve is flatter than curve A, and therefore the selectivity is less.

Curve C shows the effect of a coupling looser than critical. The energy at resonance is less than for curve A, but the curve is steeper and hence very little energy reaches the detector at non-resonant frequencies, *i.e.*, the selectivity is better than for critical coupling.

44. **Indirect Capacitive Couplings.**—A tuned secondary circuit coupled to the aerial by indirect capacitive coupling is illustrated in Fig. 33; it is also shown in another form by Fig. 21 of Section " K," and elsewhere.

The secondary circuit $L_2\,C_2$ is coupled to the aerial circuit by the condenser C'.

If the capacity in the aerial circuit is C_1, the coupling factor (Vol. I) is $K = \dfrac{C'}{\sqrt{(C' + C_1)(C' + C_2)}}$, and hence the greater the value of the coupling condenser, the greater is the coupling factor.

The variation of C' affords a very delicate adjustment of coupling, provided that L_2 is so screened from L_1 that there is no mutual inductance between them. This may be partly ensured by placing their axes at right angles to one another, but metallic screening may also be necessary.

A modification of this capacitive coupling designed to produce constant selectivity over a frequency band is seen in Fig. 29 of Section" F."

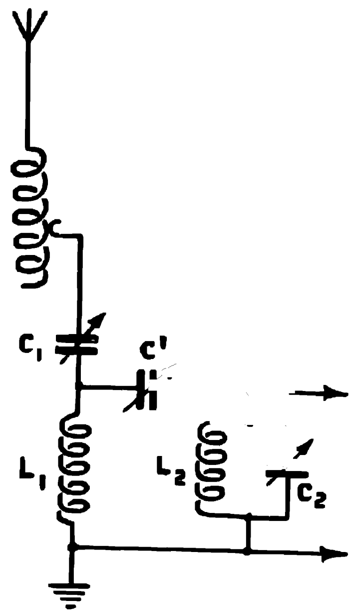


FIG. 33.

45. **Band Pass and Mixed Couplings.**—For R/T purposes, tuned circuits must not be too selective ; in order to avoid "side band cutting" (N.1, 31) it is usually necessary to pass a band of frequencies roughly 16 kc/s. in width. An effect of this nature may be obtained by utilising the double frequency effect which is always present when two tuned circuits are coupled together, using more than optimum coupling. The width of the band passed is determined by the frequency separation of the two peaks of the response curve ; when the coupling is relatively *loose* we have, approximately,

Peak separation $= Kf_0$..........(F.19).

The band width may therefore be made large or small by simply controlling the coupling factor K ; two tuned circuits, having the requisite values of Q, may be coupled to give a wide band pass circuit for R/T purposes, or **pass a narrower band of frequencies to suit the requirements of a W/T receiver.**

There are many ways in which two tuned circuits may be coupled together to make band pass effects ; Fig. 34 represents four ways of doing it by employing an aperiodic aerial and two coupled circuits. Similar effects can also be obtained with a tuned aerial and secondary circuit.

Fig. 34 (*a*) shows two tuned circuits having a **common mutual coupling,** the aerial being auto-transformer coupled to the primary. For broadcast reception, the two tuning condensers are usually ganged together. With a magnetic coupling of this nature, the peak separation increases with frequency, and therefore the selectivity *decreases*, since a wider band of frequencies is passed.

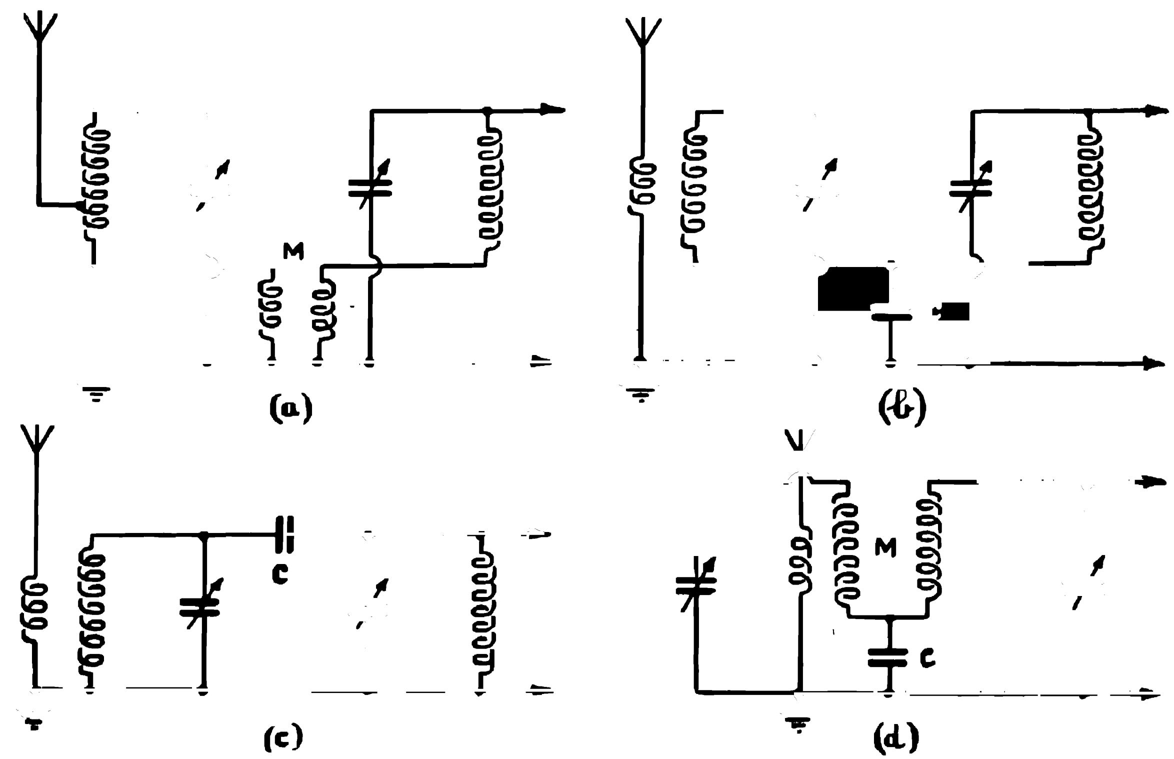


FIG. 34.

Fig. 34 (b) represents two tuned circuits coupled by a **common capacity coupling** (C), consisting of a relatively large non-inductive condenser; the two tuned circuits each *share* the condenser C, but since it is relatively large with respect to the **tuning condensers** it makes little difference to the tuning range. Since the coupling condenser C is common to each circuit, the larger the capacity the smaller will be the coupling and the greater the selectivity. Using condenser tuning, for a given capacity *C*, the percentage coupling decreases with frequency, and the selectivity *increases*, a result which is the converse of that given by Fig. 34 (*a*)

If inductance tuning were employed, and all condensers remained fixed in value, the percentage coupling would be independent of the frequency (*cf.* Vol. I for the **coupling factor**).

Fig. 34 (*c*) represents two tuned circuits coupled by **indirect capacity coupling**, sometimes called **top capacity coupling** (*cf.* Fig. 33). In this case C is a very small condenser (say 0·000005 μF); the greater the value of the capacity the "tighter" will be the coupling. With a fixed value of C, the percentage coupling increases with frequency, and the selectivity therefore *decreases*, a result which is the converse of Fig. 34 (*b*).

Fig. 34 (*d*) is an example of a "**mixed coupling.**" From the above remarks it is clear that one could devise a combination of circuits (*a*) and (*b*), or (*b*) and (*c*), which would exhibit **constant selectivity** over a given frequency range. Fig. 34 (*d*) represents a combination of (*a*) and (*b*); the increase in magnetic coupling at the high frequency end is off-set by a corresponding decrease in the capacitive coupling. Very many examples of mixed couplings have been designed, each with the same object in view. They are especially useful and necessary for the reception of R/T.

A particular practical case of top capacity coupling is seen in Fig. 29 of Section "F"; arrangements are made to obtain practically constant selectivity over a particular tuning range, by an automatic variation of the effective coupling capacity. The latter is achieved by varying the extent to which an *earthed* plate is inserted between the plates of the coupling condenser (F.39).

46. **Selectivity with Combinations of Acceptor and Rejector Circuits.**—In some of the older receiving outfits used in the Service, attempts were made to obtain the requisite selectivity by means of elaborate devices for increasing the selectivity of the aerial circuit itself. These methods are now almost obsolete, but brief reference is made to them, since the circuits illustrate important electrical principles which might, on occasions, be used in connection with some other problem.

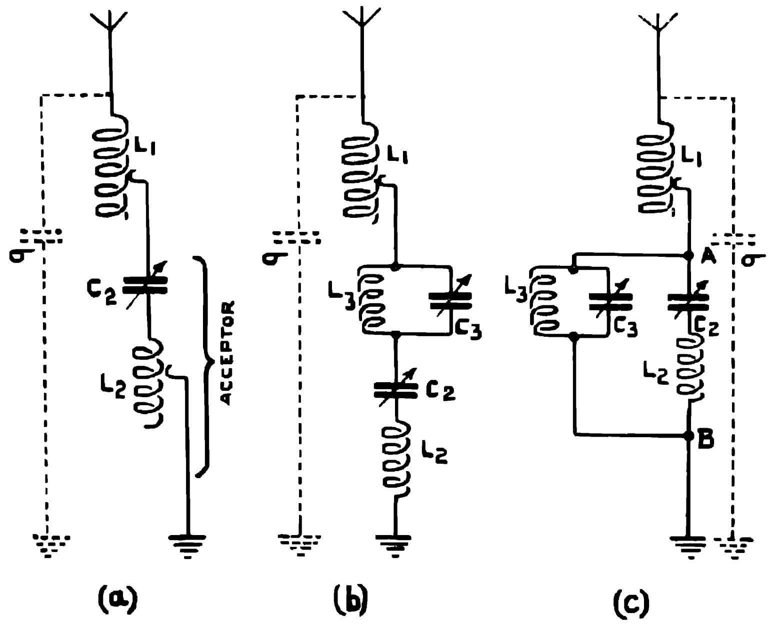


FIG. 35

The methods may be classified as follows :—

(*a*) Use of acceptor circuits in series.
(*b*) Use of rejector circuits in series and parallel.

Fig. 35 represents three of these circuits.

Fig. 35 (*a*) shows **two acceptor circuits in series,** each being separately tuned to the required frequency; in Vol. I it is shown that the combination may be made to give increased stiffness.

Fig. 35 (*b*) shows a rejector circuit joined in series with the aerial to bar some particular interfering frequency ; it is known as a **wave trap.**

Fig. 35 (*c*) is known as an **acceptor-rejector circuit,** and it consists of a rejector circuit shunted across an acceptor circuit to act as a by-pass for all interfering frequencies, both circuits being tuned to the same LC value.

47. **Receivers and Atmospherics.**—" Atmospherics " (P.18) are a source of interference universally experienced. They are the result of electric waves set up in the æther by an electric disturbance of some sort, such as flashes of lightning in a thunderstorm, and the source of origin may be at a great distance from the receiving station. They have very complex wave forms and may be directional in nature.

Various elaborate circuits have been tried for reducing them, but generally without much success, as they produce a form of shock interference of no particular frequency, and merely set the aerial in oscillation at its own natural frequency. A somewhat similar shock effect is given by the now obsolescent damped wave trains produced by spark transmitters.

Receiving aerials which have directional properties, are not so much affected by atmospherics provided that the latter arrive in a different direction from that of the signal which the aerial is adjusted to receive. In P.18, reference is made to a particular device for the elimination of atmospherics. In general, a disturbance of this nature cannot be eliminated merely by selective tuning, and the only way for the operator to deal with it is to " over read " the note of the signal he wishes to receive.

When an aerial is left in an insulated condition, a **static charge** may sometimes be picked up by the aerial and accumulated on the aerial condenser, eventually breaking down its insulation and sparking through it. To prevent accidents of this kind, an " aerial discharge coil," having a very high resistance and inductance, is frequently connected across the receiving gear to earth. On account of its high impedance only a minute current, due to the incoming signal, flows through it, but it allows static charges to leak slowly to earth.

Inductances and condensers are often protected from lightning by joining " safety points " or " **lightning arresters** " across them.

48. **Assembly of the Receiving Circuit.**—This is a highly technical matter for which it is difficult to specify fixed rules.

Ohmic losses in conductors are guarded against by providing good conductive paths wherever large currents are present. On account of " skin effect," in H/F circuits the conductors should have a good surface area. For this reason, stranded wire is used on L/F and M/F, since it presents a large area for surface currents.

All insulation must be good, especially where high voltages are present. High potential points in receivers may often be spotted by touching particular points with the finger ; in such cases signals disappear and a " plop " is heard in telephones.

All the leads should be as short, direct, and as non-inductive as possible.

Magnetic or electrostatic inter-action between components is minimised by the " lay out " of the components, and by the provision of efficient screening between them ; the problem of screening is treated more fully in Vol. I.

49. **The Telephones.**—A pair of telephones is the device more widely used for making W/T signals perceptible to the sense of hearing. Their function is to convert variations of electric current into audible signals.

For maximum power output, the impedance of the telephones at mean speech frequency should be equal to the A.C. resistance of the detector valve, or amplifying valve, delivering power to the telephones (*cf.* paragraph 42 and R.35). In the simple case, this means that the telephones must be of the type known as **high resistance telephones.** The requisite impedance matching may, however, be achieved by using **low resistance telephones** and a telephone transformer; by ordinary transformer theory, the effect of the latter is to "step-up" the effective resistance of the telephones in order to match the output impedance of the valve, or detector device (*cf.* paragraph 7).

In construction, a telephone receiver consists essentially of a permanent magnet with two pole pieces N and S on which are wound a very large number of turns of fine wire, and a diaphragm. The diaphragm is a circular piece of very thin sheet iron, supported all round the edge by the outer case or shell of the "earpiece," as close to the faces of the pole pieces as possible, without actually touching. Fig. 36 shows diagrammatically the arrangement of the component parts of a telephone.

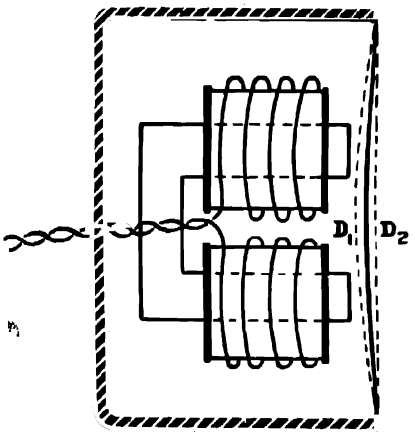


FIG. 36.

Normally, with no current passing through the coils, the permanent magnet exerts a pull on the diaphragm, which is therefore slightly bulged inwards, but not sufficiently to touch the pole pieces.

If the current is sent through the coils in such a direction that the flux due to the current assists that due to the permanent magnet, the pull on the diaphragm will be increased; it will be attracted still closer to the pole pieces, taking up a position shown by the dotted line D_1. If, on the other hand, a current passes through the coil in the opposite direction, its effect will be in opposition to that of the permanent magnet and the pull on the diaphragm will be decreased it will recede further from the pole pieces than normally, and take up position D_2. If these movements of the diaphragm are repeated at a frequency which corresponds to a note audible to the human ear, this note will be heard as a result of the wave motion set up in the air by the diaphragm vibration. Such a repetition of the diaphragm movement will occur, if an alternating current of suitable frequency is passed through the coils, or if a steady current through the coils is either increased or decreased at an audible frequency.

It can be shown that the diaphragm movement is proportional to the number of "ampere-turns" in the coils, and to the strength of the permanent magnet. To produce adequate sensitivity the ampere-turns must be high; this implies that the telephone windings must have many turns of fine wire to go into the small space of the receiver. In other words, the telephone coils must have a high inductance, in addition to having a high resistance. In practice we find that each earpiece may have a resistance as high as 4,000 ohms, or 8,000 ohms for the two joined in series. High resistance telephones are therefore also **high inductance telephones,** the inductance being the really necessary feature which helps to achieve the requisite sensitivity.

Generally, less sensitive low inductance, *i.e.*, low resistance, telephones are used in combination with a telephone transformer, the D.C. resistance of the 'phones being about 60 ohms per earpiece. Assuming a telephone transformer step-down ratio of 5 : 1, a resistance of 120 ohms in the secondary circuit becomes 3,000 ohms in the primary; two pairs of telephones in parallel would produce a reflected resistance in the primary of 1,500 ohms. From another point of view, the telephone receiver may be regarded as a current operated device; the function of the transformer is to step-

down the voltage and step-up the current. The use of low resistance telephones and a telephone transformer has the following advantages :—

(*a*) There are fewer turns in the windings, which can therefore be more robust and better insulated.

(*b*) There are lower voltages across the windings, and hence they are less likely to suffer from a breakdown of the insulation, owing to dampness or other causes.

(*c*) There are no constant current in the telephones, **the D.C. component** resulting from the detection process being eliminated by the transformer coupling. A disadvantage of a constant current is that, if the telephones are joined up the wrong way round, it will produce a flux in opposition to that of the permanent magnet, and demagnetise the latter.

(*d*) In the Service, a very important advantage of low resistance telephones is that many pairs may be joined in parallel across the secondary of a transformer. In the design stage, impedance matching is based on the estimated number of pairs of telephones to be used ; a slight variation above or below an estimated number will not seriously impair the impedance matching arrangements.

(*e*) In a ship, with much metal work, it would be dangerous to allow the H.T. for the receiver to be connected to the 'phones ; a short circuit from the winding to the telephone case would put H.T. on the operator's head.

In general, low resistance telephones produce an economy due to their durability and ow initial cost.

In a telephone of the type described by Fig. 36, the sensitivity cannot be very high, owing to the mass inertia of the diaphragm, which is necessarily considerable since it must be sufficiently strong to stand up to the pull of the permanent magnet. In an improved design, now employed in the Service, sensitivity has been increased by using a very light magnetic "moving element," and arranging the latter to actuate a delicate diaphragm of non-magnetisable material approaching the thickness of paper.

SECTION " D."

EXAMINATION QUESTIONS ON RECEPTION OF E.M. WAVES.

1. What is the action of a crystal detector when used on a receiving set? Draw a typical characteristic curve of a crystal. Why is negative grid bias usually necessary with a lower anode bend detector? Give a diagram showing how such a potential is applied.

2. Describe the action of a crystal detector. How does the response of a crystal detector vary with signals of different amplitude? Why are valves more generally used at the present time in preference to crystal detectors?

   (C. & G. Prelim., 1937.)

3. Explain the action of a crystal detector. Why is high frequency amplification before detection of particular advantage in the reception of weak signals?

   (C. & G. Prelim., 1936.)

4. Show how a valve used as a cumulative grid detector may be regarded as a combination of a diode detector and a triode amplifier. Explain why the choice of suitable values for grid condenser C and grid leak resistance R is of importance.

   (Wt. Tel. (Q.), 1937.)

5. A voltage at 100 kc/s. is applied to a cumulative grid detector, in which the effective resistance of valve and leak in parallel is 0·1 megohm. Find what the capacitance of the condenser must be in order that the P.D. applied to the grid shall be 90 per cent. of the applied voltage.

   (Result: 32·86 $\mu\mu$F. I.E.E., 1931.)

6. A small receiving valve is being used as a cumulative grid rectifier. Why will a small positive bias for the grid usually be required if the valve has a dull filament, and usually not be required if it has a bright filament? To what fundamental cause do you ascribe this difference?

   (I.E.E., 1929.)

7. If the current-potential characteristic of any device follows a "square law," show how the device may be used for rectification, and find how the signal current depends on the incoming voltage.

   (C. & G. Final, 1925.)

8.—(*a*) Draw a circuit diagram and explain the action of a triode valve used as a lower anode bend detector.

   (*b*) Comment on the statement :—" An anode bend detector may be used to give distortionless detection of a large modulated signal provided the depth of modulation is not greater than 80 per cent."

   (Qual. Lieut. (S.), 1936.)

9. Describe briefly the various methods of signal rectification available, and compare them from the point of view of :—

   (*a*) Input power. (*b*) Rectification law.

   Discuss the problem of the rectification of heterodyned signals.

   (W/T. 1 (Qual.), 1934.)

AMPLIFICATION.

RECEIVER DESIGN.

1. **The Necessity for Amplification. The Triode as Amplifier.**—The essential function of a receiver is to produce an audible result from the variations of voltage set up in a circuit coupled to the receiving aerial. In cases where the audible effect is inadequate, steps must be taken to **amplify** or increase the effect.

In general, a signal may be amplified either before or after detection, and from this have arisen the terms **R/F amplification,** and A/F amplification or **note magnification.** It should be noted that the difference between these two kinds of amplification is one of degree and not of principle; in many cases the circuits present a similar appearance.

The exceptional value of the triode as an amplifier has already been pointed out in Section "B," and it has been seen that changes in grid voltage produce a much greater effect on the anode current than the same changes in anode voltage. In order to utilise the valve as an amplifier, it is essential that these changes in anode current should produce corresponding changes in P.D. across an impedance in the anode circuit. If the changes in P.D. across the anode impedance are greater than the signal input voltages which produce them, it is clear that amplification will have been achieved, and the amplified signals may be applied between grid and filament of a succeeding valve in order to obtain still further amplification.

2. **Voltage Amplification and Power Amplification.**—A valve is a voltage operated device, and in many cases we are concerned with simple **voltage amplification.** In receivers, the R/F amplifying stages consist of valve circuits designed to make the signal oscillatory P.D. between grid and filament of the detector valve as large as possible. In some cases, however, **power amplification** (volts × amps.) is also required. Thus in the output stage of a receiver the *loudness* of the audible result depends upon the power developed in the anode impedance of the stage. At this stage, therefore, the ultimate aim in reception is to produce the maximum power expenditure in the indicating device, and voltage amplification is of subsidiary importance. It should not, however, be thought that power amplification is restricted to the A/F output stages of *receivers*. The modern "*transmitter*" usually consists of a low-power master oscillator circuit, capable of delivering a fairly constant frequency signal to an R/F power amplifier interposed between the master control circuit and the transmitting aerial.

The anode impedance must be big in relation to the A.C. resistance of the valve in order that small variations in anode current may produce large alterations in P.D. In practice, the external impedance must be at least as large as the internal impedance, and generally must be considerably larger if full advantage is to be taken of the anode current changes. From this it follows that the **dynamic characteristics** and not the static ones must be used in order to determine the behaviour of the valve and its associated circuit.

3. **R/F and A/F Amplifiers.**—Both types of amplifier have certain advantages and drawbacks, which may here be mentioned briefly.

R/F amplifiers give better results with weak modulated signals in cases where square law detectors are used. The value of the rectified current is proportional to the square of the amplitude of the applied signal voltage, and if the latter is amplified ten times before detection, the resulting rectified current will be increased a hundred times. With an A/F amplifier designed to give equal amplification, the rectified current would only be increased ten times.

Moreover, R/F amplifiers increase the selectivity of a receiver. They usually contain circuits which have to be tuned to the radio frequency being received, and therefore act as additional devices to diminish the strength of interfering signals at non-resonant frequencies. Note magnifiers are apt to amplify not only the A/F variations of voltage from the detector, but also any A/F changes in the voltage supplied to the valves, etc.

It might seem, from the above, that R/F amplification should always be used; but serious difficulties arise in the design of radio-frequency amplifiers, as compared with note magnifiers. The higher the frequency, the more difficult it is to prevent the generation of self-oscillations when

these are not wanted) by reason of energy being returned through unintentional coupling to the input circuit of the amplifier, sufficient to reduce its effective resistance to zero. This effect increases also with the amount of amplification, so that the amount of radio-frequency amplification possible is limited. If further amplification is desired, it is found convenient to carry out part of the operation before rectification, and part after it.

4. **Voltage Amplification Factor.**—Before examining the different types of amplifier used in practice, we shall investigate more exactly the behaviour of a valve with an impedance in its anode external circuit.

The Voltage Amplification Factor (usually written **V.A.F.**) of a valve and an associated impedance in its anode circuit, is defined as the **ratio of output voltage variation across this external impedance, to the input voltage variation between grid and filament of the valve.**

The external impedance may be a non-inductive resistance, an inductance or choke, a combination of these, or a parallel circuit of inductance and capacity. In addition, the anode circuit may contain the primary winding of a transformer, the voltage across the secondary of which is taken as the output voltage of the system. The effect of these impedances will now be considered. No account will be taken of the effect of inter-electrode capacities in the argument. The results obtained are thus approximate only, but represent the truth fairly well at low frequencies where inter-electrode capacitive reactance is considerably higher than the other impedances in the circuit. It should be borne in mind, however, that the instability produced by regeneration through inter-electrode capacity, and other forms of coupling between stages, is generally the factor that limits the amplification in practice. This effect will be considered later.

(*a*) **Resistance R in the Anode Circuit.**—Let us apply an oscillatory voltage variation of R.M.S. value V_g, between grid and filament of a valve, as shown in Fig. 1. (In the following theory, the notation V_g, I_a, etc., is taken as representing the **R.M.S. values** of oscillatory changes, not the steady values about which these take place.)

As a result, oscillatory variations will take place in the anode current flowing, but we cannot state that these are given by $I_a = g_m \times V_g$, where g_m is the slope of the *static* mutual characteristic.

The variations in anode current themselves cause variations in the voltage between anode and filament.

Thus, let the steady anode current be I_0 and the voltage of the high-tension battery be V_0. The actual potential difference between anode and filament is then $(V_0 - I_0R)$.

Any change in anode current I_a results in a change in anode potential given by $V_a = -I_aR$.

The general formula for the determination of I_a, as given in B.37, must therefore be used and combined with the last result to give the actual value of I_a. Thus, the two equations following are both applicable to the circuit of this "Class A" amplifier (paragraph 8) :—

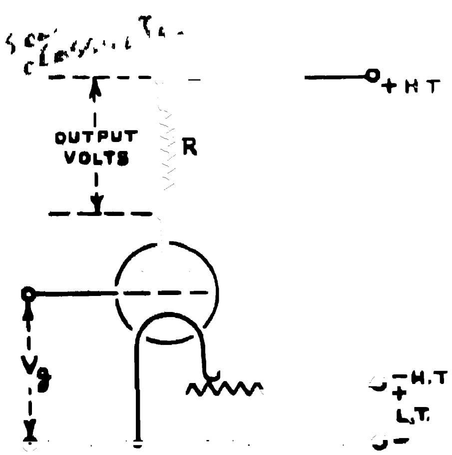


FIG. 1.

$$I_a = g_m V_g + \frac{1}{r_a} V_a$$

$$V_a = -I_aR.$$

Solving these, we obtain, by substitution for V_a in the first equation

$$I_a = g_m V_g - I_a \frac{R}{r_a}$$

$$I_a\left(1 + \frac{R}{r_a}\right) = g_m V_g$$

$$I_a = \frac{g_m}{1 + \frac{R}{r_a}} V_g \quad \dots\dots\dots\dots\dots\dots (1)$$

Also, output voltage across R

$$= I_aR = \frac{g_m R V_g}{1 + \frac{R}{r_a}}$$

Hence, the **voltage amplification factor**, which is $\frac{\text{output voltage}}{\text{input voltage}}$, is given by

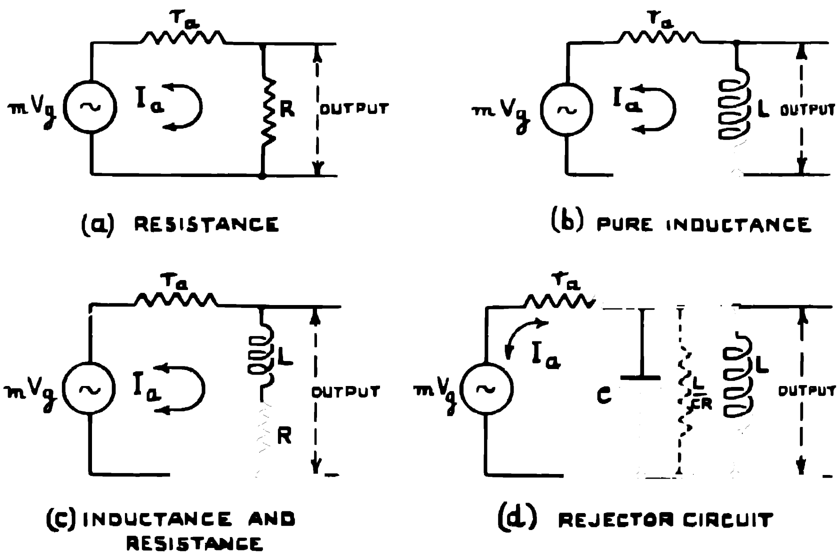
$$\frac{I_aR}{V_g} = \frac{g_mR}{1+\frac{R}{r_a}} = \frac{g_mRr_a}{R+r_a} = m\frac{R}{R+r_a} \qquad \dots\dots\dots\dots \qquad (2)$$

Equation (1) is an expression connecting the anode current variation with the grid voltage variation.

Instead of the ratio being g_m (its value on the static characteristic derived from a valve test board), it is now

$$g'_m = \frac{g_m}{1+\frac{R}{r_a}} = \frac{I_a}{V_g}$$

The quantity of g'_m is, of course, **the slope of the dynamic characteristic** with a resistance R in the anode circuit, the circuit we have taken and the consequent dependence of anode voltage on anode current being similar to that in B 39. As was stated there, the greater the external resistance the less is the slope of the dynamic as compared with the static characteristic.



SIMPLIFIED CIRCUITS FOR AMPLIFICATION

FIG. 2.

Equation (1) might be also written :—

$$I_a = \frac{g_mr_a}{R+r_a}V_g = \frac{mV_g}{R+r_a}$$

or, the R.M.S. anode current is that which would be given by a voltage equal to m times the voltage applied between grid and filament, if this voltage were applied to a circuit containing the A.C. resistance of the valve and the external resistance in series. This simplified method can be extended to the calculation of the anode current and the V.A.F. for the other types of external impedance used in amplifiers, the simplified equivalent circuits of which are shown in Fig. 2. Clearly, the equation is a form of **Ohm's law adapted to a valve circuit**; in more general form it may be written :—

$$I_a = \frac{mV_g}{Z}$$

Equation (2) shows that the V.A F. is always less than the amplification factor m of the valve, with this type of external impedance.

By increasing the value of the external resistance R relative to the A.C. resistance r_a of the valve, the V.A.F. can be made to approximate closely to m, but this course is attended by difficulties in securing adequate voltage on the anode, owing to the heavy IR drop in the resistance. This will be referred to later under Resistance-Capacity-Coupled Amplifiers.

(*b*) **Inductance L in the Anode Circuit** Fig. 2 (*b*).—Using the result given above, that the anode current change is equal to that in a simple series circuit, with m times the input voltage applied across r_a and the external impedance in series,

$$I_a = \frac{mV_g}{Z} = \frac{mV_g}{\sqrt{r_a^2 + (\omega L)^2}},$$

and the current lags behind the voltage by an angle $\tan^{-1} \dfrac{\omega L}{r_a}$.

The voltage across the inheritance L is given by

$$\omega L I_a = \frac{m \,.\, \omega L \,.\, V_g}{\sqrt{r_a^2 + (\omega L)^2}}.$$

This is the " **output voltage** " and V_g is the " **input voltage** "

Therefore the V.A.F. is $\dfrac{m\omega L}{\sqrt{r_a^2 + \omega^2 L^2}}$.

Here again, by increasing ωL, the V.A.F. can be increased almost up to m.

If the inductance L has a resistance R [Fig. 2 (*c*)], which is not negligible, the V.A.F. can easily be seen to be

$$\frac{m\sqrt{R^2 + \omega^2 L^2}}{\sqrt{(R + r_a)^2 + \omega^2 L^2}}.$$

(*c*) **Tuned Anode Circuit.**—Another form of output impedance is the " tuned anode," which consists of a tuned circuit of inductance and capacity in parallel in the anode lead Fig. 2 (*d*).

At resonance, the impedance of the circuit becomes effectively a pure resistance, given approximately by L/CR ; the latter becomes an accurate expression if the resistance R is assumed to be entirely contained in the inductive arm (Vol. I, paragraph 317).

We can, therefore, use the formula (2) obtained in part (*a*) of this paragraph, to give the V.A.F. in this case, which is

$$\text{V.A.F.} = m \frac{\dfrac{L}{CR}}{r_a + \dfrac{L}{CR}}.$$

From Vol. I we have $Z \doteqdot L/CR = Q\omega L$ or $Q/\omega C$. A large value of Q will make the rejector highly selective, by making the effective resistance at resonance relatively very large ; in this way, a voltage amplification may be obtained which is very nearly equal to m, the amplification factor of the valve.

For frequencies to which the parallel combination is not resonant, its impedance is much less than the value quoted above, so that these frequencies are not amplified to nearly the same extent, and the circuit introduces, as might be expected, an extra measure of selectivity.

(*d*) **Transformer in Anode Circuit.**—The mathematical theory of the transformer as an output circuit is more involved than in the preceding cases, and will be referred to briefly under Transformer Coupling. It may be mentioned now, that a V.A. F.greater than m can be obtained with this type of coupling.

5. **Intervalve Couplings.**—Amplifiers are generally classified according to the type of output circuit employed, the amplified voltage variations across which are applied between the grid and filament of the next valve in the model. If several **valves** are used for amplifying purposes, before or after detection, they are said to be "**in cascade.**" In such a case, the amplified variations of voltage from the first valve are further amplified by the second, and so on. The **Voltage Amplification Factor** of several valves and their associated circuits, is defined as the ratio of the output voltage across the impedance in the anode circuit of the last valve, to the input voltage across grid and filament of the first valve, and is obviously given by the product of the V.A.F.s of each valve and its external circuit taken separately.

The types of output circuit or **Intervalve Coupling** used are the same as those whose V.A.F.s are investigated in para. 4, and are generally described as follows :—

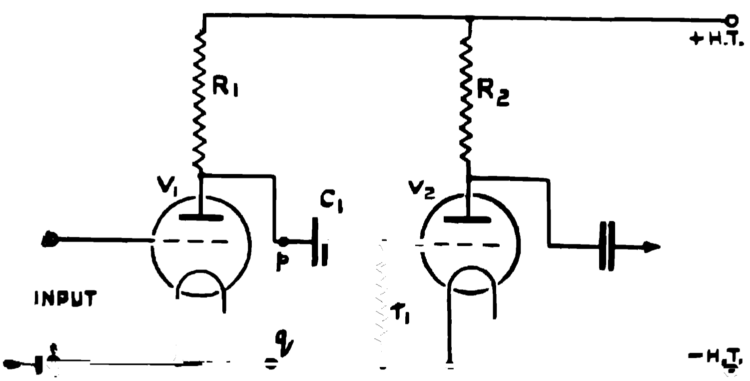
(*a*) Resistance-Capacity.
(*b*) Choke-Capacity.
(*c*) Tuned Choke-Capacity or Tuned Anode-Capacity.
(*d*) Transformer.
(*e*) Tuned Transformer.

The reason for the addition of the term "capacity" in the description will be made apparent later. With any of these, reaction may be applied, if considered necessary, to give further amplification. They will now be considered in turn.

6. **Resistance-Capacity Coupling.**—Fig. 3 represents a resistance-capacity coupled amplifier. The input voltage causes variations in anode current, which set up variations in voltage across the output circuit R_1 of greater amplitude. This output voltage has to be applied between grid and filament of the second valve, and difficulty arises with regard to the method by which this transference of voltage is to be achieved. If we join one end of the output resistance in the anode lead of valve V_1 to the grid of valve V_2, **the other end can be considered connected to the filament of valve V_2 through the H.T. battery,** and the radio frequency variations of voltage will be impressed between grid and filament of V_2.

This, however, would mean that the grid of the second valve would have a mean potential equivalent to that of the H.T. battery, whereas it should have a mean potential in the neighbourhood of that of the filament.

A condenser, C_1, is therefore inserted in the lead joining the output resistance R_1 to the grid of V_2, to insulate the grid from the steady potential of the high-tension battery. The condenser C_1 acts as a low impedance to the oscillatory voltage, but is an infinite resistance to the direct voltage of the battery.



RESISTANCE CAPACITY COUPLING

FIG. 3.

With the condenser C_1 inserted, an additional resistance r_1 is necessary between grid and filament of V_2, to serve the usual purpose of preventing an accumulation of negative charge on the grid; if the latter were left completely insulated, it would in time accumulate such a negative charge on itself that the anode current of V_2 would fall to zero.

7. **Values of Grid Condenser and Leak.**—The insulating condenser and leak have values which are determined by somewhat similar considerations to those in D.20, where their use in detector was investigated.

In the case of an amplifier, the essential feature of the apparatus is to pass on to the second valve a waveform of voltage which is the same as that received by the first valve, so that it is necessary to minimise, as much as possible, the typical action of a condenser and leak, which is to introduce asymmetry into the waveform, if grid current flows. There is therefore no definite reason for having a small value of C_1, the grid condenser, as was shown to be advisable for cumulative grid detection in D.20, because here we want to prevent variation in the mean potential of the grid.

On the other hand, the statements of "D" section, on the value of the grid leak, are very relevant to the case of the amplifier.

The amplified output voltage, which is available between points p and q, in Fig. 4 (a), is impressed across a circuit which consists of the condenser C_1, and a parallel combination between grid and filament of the *next valve*, made up of the leak resistance r_1, the grid-filament capacity C_{gf} of the valve, and the effective resistance of the valve between grid and filament, R_{gf}.

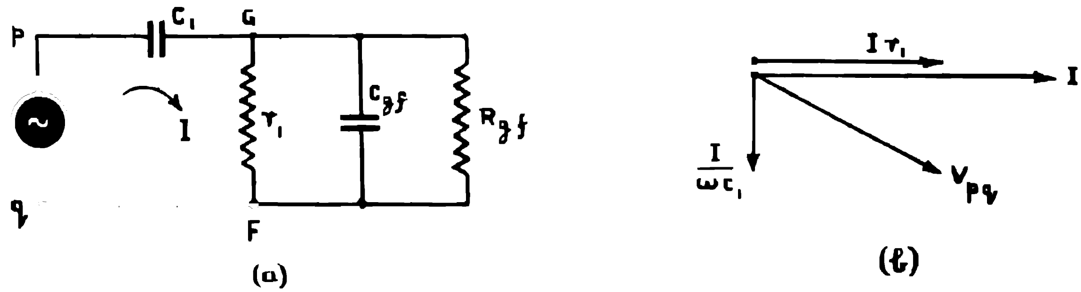


FIG. 4.

We want the voltage across GF, which is that actually applied between grid and filament, to be as large a fraction of that across pq as possible. Hence, the potential fall across C_1 should be negligible, or C_1 should have a small reactance compared with the parallel circuit GF. This means that the capacity of C_1 should be large. If the circuit attached to GF consisted only of r_1, the matter could be simply represented by the vectors of Fig. 4 (b). The larger one makes C_1 the smaller will be the vector $I/\omega C_1$; the vector V_{pq} will swing round towards the I vector, and there will be little difference in length between it and the vector Ir_1.

Now consider the parallel circuit. The effective resistance, R_{gf}, is usually of the order of a megohm, and the reactance of C_{gf} is also high at low frequencies, because of the small value of C_{gf}. The leak, r_1, should therefore be of the same order or larger, so that the total effective impedance between G and F is not sensibly diminished by a low value of leak. At higher frequencies, the reactance of C_{gf} is the limiting factor.

The conditions are therefore similar to those in cumulative grid detection, except that there is no definite reason for having C_1 small. We obtain the general result, that **in amplifiers the condenser has a larger value than in detectors, while the leak is generally somewhat smaller**; this is to allow the accumulated electron charge on the condenser, which cannot entirely be avoided unless by grid bias (*see* paragraph 8), to leak away in a time depending on CR.

The only practical limit to the increase in value of C_1 is that, if it is made too large, its reactance becomes small for stray audio-frequency voltages in the amplifier, and these are passed on sensibly undiminished, in the same way as the radio frequency variations.

The coupling and insulating condensers used in note magnifiers have, of course, to be much larger than those used in radio frequency amplifiers. In Service R/F amplifiers the value of R ranges from 10^5 to 10^6 ohms, that of C varying from 0·1 jar to 1 jar respectively. In A/F amplifying stages R varies from 2×10^5 ohms to 2×10^6 ohms, the corresponding value for C ranging from 90 jars to 10 jars.

8. **Use of Negative Grid Bias. Distortionless Class "A" Amplification.**—If the mean grid potential on a valve is such that the oscillatory voltage input gives instantaneous values of grid potential corresponding to points on the curved regions of the dynamic mutual characteristic, rectification takes place (Section "D"). The anode current waveform, and therefore the output voltage waveform, is not a replica of the input voltage waveform, *i.e.*, distortion has been produced. The first requisite for distortionless amplification is, therefore, that the range within which the grid voltage varies, owing to an oscillatory input voltage, should be confined to the straight part of the characteristic; this defines what is now usually known as Class "A" amplification (*cf.* K 11).

Distortion is also caused if grid current flows during any part of the input voltage cycle, for the damping of the input circuit, and hence the oscillatory grid voltage, is thereby caused to vary. To prevent this, the grid may be kept at a steady negative potential (bias), large enough to prevent the peak grid voltage during the positive half cycle of the input from extending into the region where grid current flows.

If the grid is biased negatively to a point half-way (**mid point biasing**) between the voltage at which grid current starts to flow and the voltage at which the dynamic characteristic curve starts to bend, the best condition is achieved; provided the amplitude of the grid swing is less than the amount of bias, distortionless amplification will then be secured.

In a multi-valve amplifier, the grid swing increases from valve to valve, so that longer straight portions of the mutual characteristic curve on the negative side of zero grid volts are necessary in the later stages of amplification, especially in note magnifiers; this is secured by having valves of low amplification factor m, with the inherent disadvantage that the V.A.F. is thereby diminished. This is unimportant in the last stage, where *power amplification* is required, provided g_m is high.

9. **H.T. Voltage with Resistance Coupling. Resistance-Capacity Coupling** has one great disadvantage compared with other types of coupling.

We saw, in paragraph 4, that the V.A.F. is given by $m\dfrac{R}{R + r_a}$, and consequently, by increasing R, the V.A.F. can be increased.

With this type of coupling, however, it is essential to take into consideration the steady component of anode current which passes through the external resistance, as well as the oscillatory component.

The mean anode voltage is given by the voltage of the H.T. battery, minus the I_0R drop in the external resistance, where I_0 represents the mean value of anode current. As R is increased, this I_0R drop increases, and the voltage on the anode diminishes.

A fairly high anode voltage is necessary if the straight part of the mutual characteristic is to extend over the working region for distortionless amplification, and, even if distortion is accepted, it is still the case that the A.C. resistance of the valve increases as the working point approaches the lower bend, and so the V.A.F. of the stage is diminished. Hence, both for quality and sensitivity, the anode voltage should be high.

In practice, this is achieved by increasing the H.T. voltage, and, as there are practical limits to the extent to which this can be done, it is seldom that the value of R is greater than three or

four times the value of the A.C. resistance of the valve. Taking the typical figure of $R = 3r_a$, the V.A.F. $= \frac{3}{4}m$.

10. **Numerical Examples on R/C Coupling.**—(*a*) In a resistance-coupled amplifier (one valve) there is a V.A.F. of 4 with a resistance of 20,000 ohms in the anode lead. With the resistance changed to 40,000 ohms, the V.A.F. is 5. Find the mutual conductance g_m of the valve.

$$\text{V.A.F.} = m\frac{R}{R + r_a},$$

so that the following equations hold .—

$$4 = \frac{20{,}000m}{20{,}000 + r_a}$$

$$5 = \frac{40{,}000m}{40{,}000 + r_a}$$

$$\therefore \quad 80{,}000 + 4r_a = 20{,}000m.$$
$$\text{and } 200{,}000 + 5r_a = 40{,}000m.$$

Multiplying the first by 2, we obtain

$$160{,}000 + 8r_a = 40{,}000m.$$
$$40{,}000 - 3r_a = 0.$$
$$\therefore \quad r_a = 13{,}333 \cdot 3 \text{ ohms.}$$

Also

$$80{,}000 + 4\,(13{,}333 \cdot 3) = 20{,}000m.$$
$$133{,}333 \cdot 3 = 20{,}000m$$
$$\therefore \quad m = 6 \cdot 7.$$

$$\therefore \quad g_m = \frac{m}{r_a} = \frac{6 \cdot 7}{13{,}333 \cdot 3 \text{ ohms}} = \frac{1}{2{,}000 \text{ ohms}}$$
$$= \frac{1 \text{ amp.}}{2{,}000 \text{ volt}} = 0 \cdot 5\,\frac{mA}{\text{volt.}}$$

(*b*) The static mutual characteristic curve of a valve, for $V_a = 50$ volts, can be represented as a straight line from zero value of I_a at $V_g = -1$ volt to a value of $I_a = 1mA$ at $V_g = +3$ volts before becoming horizontal.

The amplification factor (m) of the valve is 8, and grid current starts flowing as soon as V_g is positive.

The valve is to be used in a resistance coupled amplifier, giving a V.A.F. $= 6$. Find the voltage of the H.T. battery, if it is desired to obtain distortionless amplification with an oscillatory grid voltage of amplitude 2 volts and a mean current of $0 \cdot 5\ mA$ flowing through the valve.

The first step is to find the external resistance, and hence the voltage drop across it.

$$\text{V.A.F.} = \frac{mR}{R + r_a}.$$

$$\therefore 6 = \frac{8R}{R + r_a}.$$

$$g_m = \frac{1mA}{4 \text{ volt}} = \frac{1}{4{,}000}\frac{\text{amp.}}{\text{volt}}.$$

$$\therefore \quad r_a = \frac{m}{g_m} = 8 \times 4{,}000\,\frac{\text{volt}}{\text{amp.}} = 32{,}000\ \Omega.$$

$$\therefore \quad 6R + 6 \times 32{,}000 = 8R,$$
$$2R = 6 \times 32{,}000,$$
$$\therefore \quad R = 96{,}000\ \Omega$$

Voltage drop in $R = 96{,}000 \times 0 \cdot 5mA =$ **48 volts.**

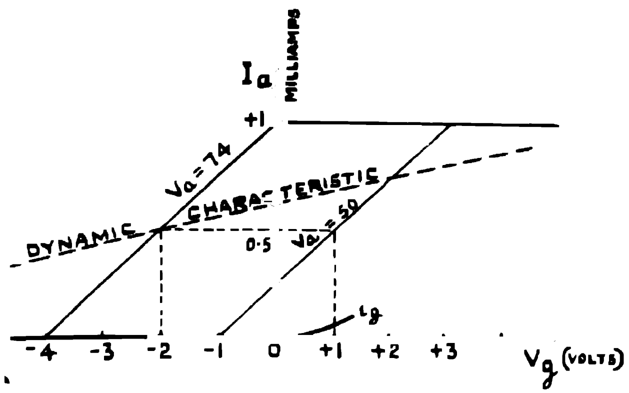


FIG. 5.

The steady grid voltage V_g corresponding to an anode current of $0 \cdot 5 mA$ must be $-2 \cdot 0$ volts to prevent grid current flowing. On the given characteristic, $0 \cdot 5 mA$ corresponds to $+1$ grid volt, so the characteristic has to be shifted to the left through three grid volts. As we are dealing with static characteristics, this corresponds to an increase of $3m = 24$ volts on the anode. The anode voltage required is, therefore, $50 + 24 = 74$ volts, and the H.T voltage required is $74 + 48 = 122$ volts.

A portion of the dynamic characteristic is indicated in the figure.

11. **Resistance-Capacity Amplifiers. General Remarks.**—Resistance-capacity coupling is often used in **A/F amplifying stages,** but is less frequently employed in R/F amplifiers. The amplifier as used in A/F work is essentially the same as that used for R/F work, the main difference being that the coupling condenser must be larger (paragraph 7) ; it possesses the great advantage that it gives **equal amplification** over the full range for which the amplifier is designed, the V.A.F. being independent of the frequency. It is for the latter reason that this coupling finds application in high fidelity A/F amplifying stages (Appendix A.8) ; we shall see that some other types of coupling give varying results in this respect.

No tuning is necessary to give best results, but if used in R/F amplifying stages, the absence of tuned circuits means, however, that no further selectivity is introduced in the receiving model by its use. There is the further disadvantage that higher anode voltages than usual are necessary to compensate for the voltage drop in the external circuit. This is avoided in all other types of amplifier coupling.

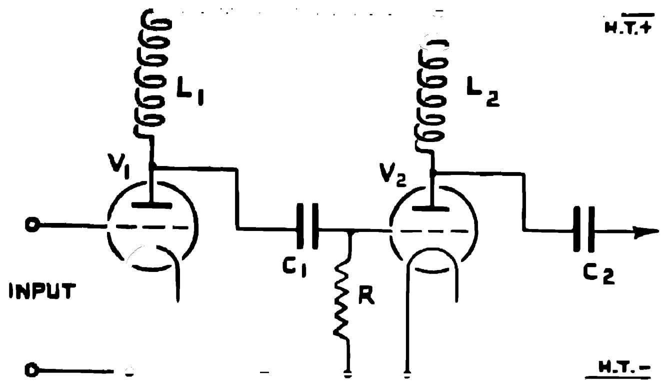
Resistance-capacity amplifiers are **unsuitable for high radio-frequencies,** since the resistances used have self-capacities in parallel with themselves, the value of the shunt impedance which they provide being lower the higher the frequency. Moreover, the anode-filament capacity of the valve, and the grid-filament capacity of the succeeding valve, together form a low impedance in parallel with the resistance ; the effective value of the external impedance becomes lower the higher the frequency, and a smaller V.A.F. is obtained. These shunt impedances constitute a factor limiting the value of the resistance-capacity amplifier mainly to A/F stages ; the effective impedance at 1,000 kc/s. is often as low as 4,000 ohms. Several artifices have been tried in order to keep the external impedance at a high value as the frequency is raised ; one of these consists in including a small inductance in series with the resistance.

The most serious argument against the use of this coupling in R/F stages is provided by the results of Miller effect (N.29). The anode circuit has mainly capacitive reactance, and, acting through the anode-grid inter-electrode capacity, constitutes a load damping the input circuit ; this has an adverse effect on the selectivity.

From the foregoing account it will be appreciated that the oscillatory variations of anode current and voltage are superimposed upon their steady pre-signal values ; under certain conditions, therefore, the anode oscillatory voltage V_a could vary from zero to twice its pre-signal value, given by $V_0 - I_0 R$ (paragraph 4, and *cf.* Fig. 4 of K.6).

12. **Choke-Capacity Coupling.**—Fig. 6 represents a choke-capacity coupled amplifier.

The general arrangements of the circuit are similar to those for resistance-capacity coupling. The amplified voltage variations across the output circuit (in this case the inductance L_1) are applied between grid and filament of the next valve and so on.



CHOKE CAPACITY COUPLING.

FIG. 6.

For the same reason, the insulation of the grid valve V_2 from the direct voltage of the H.T. battery, a coupling and insulating condenser C_1 is inserted; this makes it necessary to complete the circuit from grid to filament by a leak resistance R, to prevent an accumulation of negative charge on the grid.

The grid may, of course, be biased negatively so as to ensure that there is no flow of grid current even when the oscillatory potential between grid and filament has its maximum positive value.

The values of the condenser and leak are determined by considerations similar to those for resistance coupling.

It is unnecessary to have a much higher voltage in the H.T. battery than is actually required between anode and filament, because the resistance of the choke may be considered small, and hence the IR drop across it due to the steady component of anode current is small.

This is in no way contradictory to the statement that the reactance of the choke to the oscillatory component of anode current may be very large.

13. **V.A.F. with Choke-Capacity Coupling.**—We saw in paragraph 4 that the V.A.F. is given by :—

$$\text{V.A.F.} = \frac{m\omega L}{\sqrt{r_a^2 + \omega^2 L^2}}$$

By increasing the value of "ωL," we can make the V.A.F. approximate closely to m, without the previous difficulty as regards H.T. supply.

The following conclusions can be drawn from the formula :—

(1) The chokes used in note magnifiers must be very much larger than those used in radio-frequency amplifiers, to give corresponding values of amplification. In practice, in note magnifiers, iron-cored chokes are used, whose inductance is of the order of henries.

For example, with an audio-frequency of 500, ω is about 3,000, and a choke of about 20 henries would be necessary to give a reactance of 60,000 ohms.

(2) For equal values of external reactance and resistance the choke amplifier gives better amplification than the resistance amplifier.

If we write $R = \omega L = x \times r_a$, the V.A.F. for the choke amplifier may be written in the form $\dfrac{mx}{\sqrt{1 + x^2}}$ and for the resistance amplifier in the form $\dfrac{mx}{1 + x}$

If $x = 1$, the first result is $\dfrac{m}{\sqrt{2}} = 0{\cdot}71\, m$, and the second $\dfrac{m}{2} = 0{\cdot}5\, m$.

If $x = 3$, the first result is $\dfrac{3m}{\sqrt{10}} = 0{\cdot}95\, m$, and the second $\dfrac{3m}{4} = 0{\cdot}75\, m$.

(3) **The amplification secured from a given choke varies with the frequency amplified.**—This is a great disadvantage with this type of amplifier, and the unequal results achieved are made even more pronounced by the possibility that the inductance may form a resonant parallel circuit with the inherent self-capacity of the coil, and give exceptionally good amplification for one particular frequency as compared with all others. This point is discussed in the next paragraph.

14. **Peaky Amplification.**—If the self-capacity, as above, combines with the inductance to form a resonant parallel circuit for one particular frequency included in the range of the amplifier, it is no longer correct to assume that the reactance of the choke is ωL. Actually the impedance of the circuit now becomes a pure resistance, whose effective value is given by $\frac{L}{CR}$, where R is the ohmic resistance of the coil.

If R is low, as is generally the case, this value is probably very much greater than the value of "ωL," calculated for other frequencies, within the range of the amplifier, lower than and considerably removed from the resonant case. This results in specially good amplification at the particular resonant frequency. This effect is known as "**Peak Effect**" and is an undesirable feature in an amplifier.

Generally, therefore, steps are taken to destroy the peak effect, or to utilise it by arranging for it to occur at some desired frequency.

It may effectively be **avoided** by winding the choke with high-resistance wire, which reduces its impedance at resonance, and, in addition, tends to equalise the amplification at non-resonant frequencies, because of the introduction of a resistance term into the expression for V.A.F.

The full expression (paragraph 4) for the V.A.F. is

$$\text{V.A.F.} = m\frac{\sqrt{\omega^2L^2 + R^2}}{\sqrt{\omega^2L^2 + (R + r_a)^2}}.$$

If R is comparable with ωL, the variation in V.A.F. for different values of ω is not nearly so great.

It is better, however, to try **to take advantage of the peak effect** for different frequencies, and this can be done in two ways. Obviously, it is necessary to secure the resonant parallel circuit condition.

(*a*) The inductance may be large enough, in combination with its total self-capacity, to be resonant to a wave-length longer than, or a frequency smaller than, the limiting value for which the amplifier is designed.

If the inductance is then provided with several tapping points, it is possible to get a rough approximation to resonance at any desired frequency using a portion of the inductance and the corresponding self-capacity

(*b*) The inductance may be so small that, in combination with its total self-capacity, the resultant LC value is less than that necessary for resonance at any of the frequencies passing through the amplifier. If, in this case, we join an artificial condenser across the inductance, whose value can be varied, it will be possible to make the parallel circuit exactly resonant to any desired frequency, the condenser capacity being in parallel and therefore additive to the self-capacity of the choke.

This method of coupling is known as Tuned-Choke or Tuned-Anode Coupling, and will be discussed in the following paragraph.

15. **Tuned-Anode Coupling.**—Fig. 7 represents an intervalve coupling in which the addition of the condenser C_1 across the anode choke L_1 makes it possible for the circuit L_1 C_1 to be tuned to resonance at any desired frequency.

If the ohmic resistance of the circuit is R, its effective resistance is $\frac{L_1}{C_1R}$, which may be very large indeed.

The V.A.F. is therefore (paragraph 4) equal to

$$m\frac{\frac{L_1}{C_1R}}{r_a + \frac{L_1}{C_1R}}$$

The **advantages** of this type of output circuit are :—

(1) By increasing the stiffness of the circuit, *i.e.*, by having a large value of L on a small value of C, and by using low resistance wire, the V.A.F. can be made very nearly equal to m. In practice, a limit is set to the decrease of C, because it consists partly of the unavoidable self-capacity of the choke, and the inter-electrode capacities of the valve, in parallel with the artificial condenser provided for tuning purposes. Further, the design of a tuned anode stage is ruled more by considerations of stability and selectivity than of the maximum possible amplification.

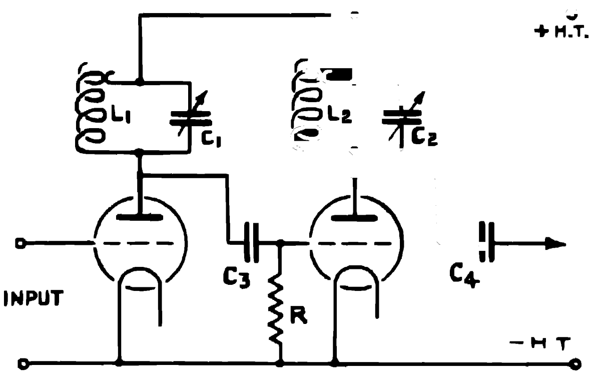


FIG. 7.

(2) At the same time, the circuit is free from the disadvantage attendant on pure resistance coupling as regards H.T. supply, because the steady component of anode current flows through the inductive arm of the parallel circuit, whose ohmic resistance is small, and hence there is no big IR drop to be compensated for by extra high-tension voltage.

(3) **A high measure of selectivity is achieved.**

For interfering frequencies, to which the circuit is definitely mistuned, the impedance of the circuit becomes, instead of a high resistance, a reactance (inductive or capacitive) of considerably less value. If the interfering frequency is less than the resonant frequency, the reactance is inductive ; if greater, it is capacitive. The further the interfering frequency is from resonance, the less is the value of the reactance. A very high measure of amplification is thus obtained over a narrow band of frequencies in the vicinity of resonance.

With several stages of amplification of this type, the selectivity can be increased to a very great degree.

The **disadvantages** of this type of coupling are :—

(1) To ensure the marked selectivity mentioned above, various tuned circuits have to be adjusted accurately to the incoming frequency, which means longer time spent in adjustment and difficulty in moving from one frequency to another.

(2) The tuned circuits involved in this type of amplifier make it very difficult to prevent the generation of self-oscillations.

This subject is treated further in paragraph 34.

Tuned-Choke Capacity Coupling used in note magnifiers is similar to that in radio frequency amplifiers, except that the values of inductance and capacity are much larger, so that the circuit may be resonant to audio-frequencies. Coupling condensers are necessary, as before, their values being much higher for note magnifiers than for radio frequency amplifiers.

16. Numerical Examples on Choke and Tuned-Anode Coupling.

Example (a)

Find the V.A.F. of a choke amplifier wound with high resistance wire, L = 7,500 mics., R = 20,000 ohms.

The valve has r_a = 16,000 ohms, and m = 10. The applied frequency is such that $\omega = 2 \times 10^6$.

With the above data, find also the R.M.S. value of I_a in micro-amperes, when a voltage of 0·1 volt (R.M.S.) is applied between grid and filament.

(i)

$$\text{V.A.F.} = m \frac{\sqrt{R^2 + (\omega L)^2}}{\sqrt{(R + r_a)^2 + (\omega L)^2}}$$

$$\omega L = \frac{7{,}500}{10^6} \times 2 \times 10^6 = 15{,}000 \text{ ohms.}$$

$$\therefore \quad \text{V.A.F.} = 10 \frac{\sqrt{(20{,}000)^2 + (15{,}000)^2}}{\sqrt{(36{,}000)^2 + (15{,}000)^2}} = 10 \frac{\sqrt{20^2 + 15^2}}{\sqrt{36^2 + 15^2}}$$

$$= \frac{10 \times 25}{39} = \frac{250}{39} = 6 \cdot 41.$$

(ii)

$$I_a = \frac{mV_g}{\sqrt{(R + r_a)^2 + (\omega L)^2}} = \frac{10 \times 0 \cdot 1}{\sqrt{(36{,}000)^2 + (15{,}000)^2}} \text{ amperes}$$

$$= \frac{10^6}{\sqrt{(36{,}000)^2 + (15{,}000)^2}} \mu A = \frac{10^6}{39{,}000} \mu A$$

$$= 25 \cdot 64 \, \mu A.$$

Example (b).

In a tuned-anode amplifier the coil has L = 360 mics., and its resistance is 6 ohms, C being 0·5 jar.

A signal voltage of 0·15 volt at resonant frequency is applied to the grid of the valve, which has $m = 8$ and $g_m = 0 \cdot 4 \frac{mA}{volt}$. Ten per cent. of the stepped-up voltage is lost in the coupling condenser to the next valve. What is the voltage available between grid and filament of the next valve?

$$\frac{L}{CR} = \frac{\frac{360}{10^6}}{\frac{0 \cdot 5}{9 \times 10^8} \times 6} \text{ ohms} = \frac{360 \times 9 \times 10^8}{3 \times 10^6} = 108{,}000 \text{ ohms.}$$

$$r_a = \frac{m}{g_m} = \frac{8}{0 \cdot 4 \frac{mA}{volt}} = 20{,}000 \text{ ohms.}$$

$$\text{V.A.F.} = 8 \times \frac{108{,}000}{108{,}000 + 20{,}000} = \frac{8 \times 108}{128} = 6 \cdot 75$$

$\therefore$ voltage applied to the next valve is

$$\left(6 \cdot 75 \times 0 \cdot 15 \times \frac{9}{10}\right) \text{ volts} = 0 \cdot 91 \text{ volt.}$$

The selectivity of the tuned anode amplifier can be illustrated numerically.

The resonant frequency of the circuit above is very nearly equal to $\frac{3 \times 10^4}{2\pi\sqrt{LC}}$ kc/s.

Substituting for L and C, $f = \frac{3 \times 10^4}{2\pi\sqrt{180}}$ kc/s.

$$= 356 \text{ kc/s.}$$

Let us find the impedance of the circuit for incoming frequencies of 100 kc/s. and 500 kc/s. respectively.

In the former case, the frequency is so much below the resonant frequency that the impedance will be sensibly the reactance of the inductance alone.

$$\omega L = 2\pi \times 100 \times 1{,}000 \times \frac{360}{10^6} = 72\pi = 226 \text{ ohms.}$$

$$\frac{1}{\omega C} = \frac{9 \times 10^8}{2\pi \times 100 \times 1{,}000 \times 0{\cdot}5} = \frac{9{,}000}{\pi} = 2{,}845 \text{ ohms.}$$

The parallel combination (neglecting resistance) gives a make-up current $I = V\left(\frac{1}{\omega L} \sim \omega C\right)$ and therefore a reactance of $\frac{\omega L}{1 - \omega^2 LC}$

At 100 kc/s, this reactance $= \dfrac{226}{1 - \dfrac{226}{2{,}845}} = 245$ ohms, practically that of the inductance alone.

This reactance is so low compared with the impedance of the valve that no amplification is secured.

In the same way, with a frequency considerably above that of resonance, the reactance of the parallel circuit is effectively that of the capacity.

For a frequency of 500 kc/s, $\frac{1}{\omega C} = 569$ ohms, again so low that no amplification is achieved.

17. **Transformer Coupling and Tuned Transformer Coupling.**—Fig. 8 shows a type of intervalve coupling in which the primary of a transformer is included in the anode circuit of the first valve, and the voltage variations induced across the secondary are applied to the grid of the next valve.

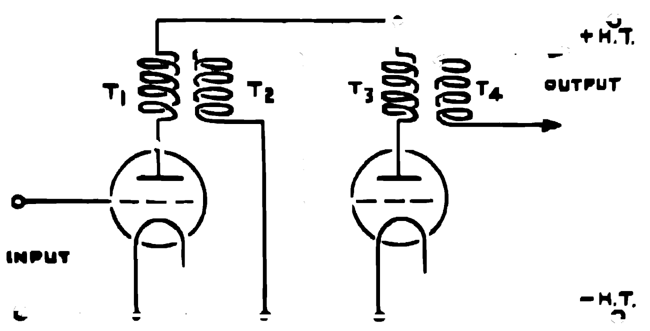


FIG. 8.

The essential difference between the transformer circuit and those previously dealt with, is that no coupling and insulating condenser and grid leak are necessary, because there is no direct connection from the anode of one valve to the grid of the next valve.

If a condenser and leak are found in the secondary circuit of a transformer coupling, it is a definite indication that the next valve is a detector valve, using cumulative grid detection.

Now, in the transformer, as in the case of choke coupling, there are self-capacities between the turns of the primary and also of the secondary, and similar results, as regards "peaky" amplification, may be expected.

The same steps can be taken either to eliminate the peak effect, or to utilise the exceptionally good amplification it gives by tuning either the primary or secondary inductances, or both, so that they form resonant circuits to the received frequency.

Thus, to avoid peak effect, resonance effects may be annulled by winding the transformers with high-resistance wire.

To utilise the peak effect, either

(*a*) the primary or secondary windings, or both, may be provided with several tapping points, so that the inductance included in the circuit, together with its self-capacity, may form a roughly tuned circuit to the frequency being amplified ; or

(*b*) a variable condenser may be joined across either, or both, windings of the transformer.

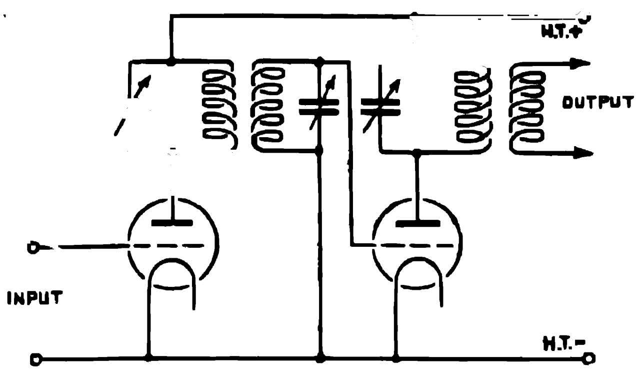


FIG. 9.

With several transformer couplings in the model, these modifications make tuning a slow process, unless it can be arranged for one handle to adjust the condensers, or the tappings, simultaneously.

Selectivity is, of course, greatly increased.

It should be noted that the above remarks, and Fig. 9 in particular, apply specifically to R/F transformer coupling ; in that case, it is easy to arrange for the requisite small values of coupling factor K to produce overall response curves having a single peak [*cf.* paragraph 19 (*e*)]. By the ordinary theory of coupled tuned circuits, over-coupling produces a double-hump effect which may be utilised when band-pass features are desirable.

The more tightly coupled A/F transformer can seldom employ tuned circuits, but in certain cases of " note selectors " a tuned secondary is employed (*cf.* paragraph 39).

18. **Transformer Step-up, and Construction.**—The transformers used in radio-frequency amplifiers and those used in note magnifiers differ in certain respects.

In **note magnifiers,** the transformers are wound on an iron core, which gives the result that the magnetic leakage is negligible, or the coupling is approximately perfect (100 per cent.).

In addition, the ratio of secondary to primary turns may be as much as 5 to 1, or 7 to 1, giving a bigger voltage variation across the secondary than the primary circuit, and hence a high voltage amplification factor.

In **radio-frequency amplifiers,** the transformers are wound on a former of insulating material, with as little self-capacity between turns as possible. This incurs the disadvantage that there is considerable leakage, and the coupling is very much looser than if iron were used. The usual iron core would be unsuitable, on account of the excessive eddy current and hysteresis losses which would occur at the high frequencies involved ; it is, however, possible to achieve satisfactory results using the new powdered iron-cores which have been developed, and are frequently used in conjunction with **permeability tuning** (paragraph 20).

In general, it is found that very little advantage is gained by increasing the number of secondary turns above the number of primary turns, owing to the effects which arise due to the self-capacities of the coils and the valve inter-electrode capacities.

★19. **Transformer Coupling Theory.**—The theory of the V.A.F. obtainable with transformer coupling is complicated, and only an introduction to it will be given here. It should further be noted that the type of amplification required depends on whether it is intended for W/T or R/T signals ; for W/T signals amplification is desired at particular frequencies, but for R/T the essential requirement is uniform amplification over a band of frequencies (*cf.* N.32, 63). Most of the remarks below apply to W/T reception only.

In simple transformer theory (Vol. I) it has been seen that the performance depends very much on a number of factors, among which are the nature of the secondary load, the degree of coupling, and the relative value of the inductive reactance of the primary with reference to the **reflected secondary impedance.**

Certain special cases are here briefly discussed.

(a) **A/F TRANSFORMER COUPLING.**—We will assume an ideal case, in which the leakage resistance and reactance are zero, the coupling factor being 100 per cent.

If the grid is separately biased so that no grid current flows, the secondary circuit is closed in the inter-electrode capacity; if grid current is flowing, the secondary circuit is closed by a resistance which decreases in value as the grid of the valve is made more positive to the filament. It is generally assumed that, even with a large negative bias, the effective slope resistance between grid and filament, which we may call R_g, is never infinite, and has, in fact, a value of the order of 1 to 10 megohms.

In the general treatment of transformer theory outlined in Vol. I, paragraph 348 (*b*), it is shown that the reflected load impedance could be regarded as acting in parallel with the primary inductance of the transformer. For that reason Fig. 10 (*b*) represents the valve equivalent circuit, where the secondary load is a pure non-inductive resistance R_g, the grid-filament effective resistance of the next valve.

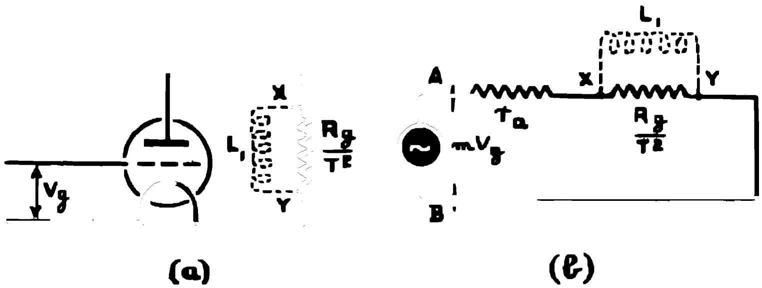


FIG. 10.

If ωL_1 is large and considerably greater than R_g/T^2, the magnetising current approaches zero and the effective impedance between the points XY is largely controlled by the value of R_g/T^2, and the circuit may be simplified by eliminating the inductive arm L_1. In that case, as in paragraph 4, the anode current variation is the same as if "m" times the input voltage were applied across a circuit consisting of r_a and R_g/T^2 in series, *i.e.*,

$$I_a = \frac{mV_g}{r_a + \frac{R_g}{T^2}}.$$

The voltage variation across $\frac{R_g}{T^2}$, in the primary circuit, is given by—

$$I_a\frac{R_g}{T^2} = \frac{m\frac{R_g}{T^2}V_g}{r_a + \frac{R_g}{T^2}}$$

The output voltage across the secondary is T times this, and is therefore

$$V'_g = \text{output volts} = \frac{m\frac{R_g}{T}\cdot V_g}{r_a + \frac{R_g}{T^2}}.$$

$$\therefore \quad \frac{V'_g}{V_g} = \frac{\text{output volts}}{\text{input volts}} = \frac{m\frac{R_g}{T}}{r_a + \frac{R_g}{T^2}}.$$

If $\frac{R_g}{r_a}$ is written a, this is equivalent to $\frac{Tma}{T^2 + a}$.

The V.A.F. therefore varies directly as m, as might be expected, and also depends in a complex manner on T and a.

It increases with an increase in "a," and hence it is desirable to make the grid-filament slope resistance of the next valve as high as possible by biasing the grid. Theoretically, the V.A.F. approaches the limiting value mT as "a" approaches infinity. Alternatively, the case of the secondary on open circuit may be considered as shown, later, below.

As regards the dependence of the V.A.F. on T, it is easy to prove that, with m and a constant, the above expression has a maximum value when $T = \sqrt{a}$, and the V.A.F. then $= \frac{mT}{2} = \frac{m\sqrt{a}}{2}$. The values of R_g and r_a found in practice, and the resulting ratio a are such that greatest amplification occurs when T is not more than 5 or 6, and so furnish a theoretical reason for the statements of paragraph 18. This formula only applies for low frequencies. [*Cf.* (*f*) below.]

The practical audio-frequency transformers differs from the ideal case so far considered. We may still assume that the leakage reactance and resistance are negligible, but the magnetising current can only be small if the no-load reactance of the primary is very large. This means that **the simple circuit assumed above should be modified by having the primary reactance in parallel with** R_g/T^2.

The effect of this is to diminish the V.A.F. below the value originally obtained, although the difference is not great if the reactance is large compared with R_g/T^2. The only limit to this is imposed by the fact that a large primary requires an even larger secondary, and beyond a certain value the leakage losses and the self capacities of the windings would result in a decrease in amplification.

In all cases we have

$$\text{V.A.F.} = \left(\frac{Z_{XY}}{Z_{AB}}\right)mT$$

.... where Z_{XY} and Z_{AB} represent the impedances between those points respectively.

Case " f " (below) illustrates the procedure in a case where it is necessary to take into account the value of the primary inductance.

If the secondary is on open circuit, the secondary resistance may be regarded as infinite, and the reflected load R_g/T^2 will be infinite too. In that case the effective impedance between points X and Y is that of the primary of the transformer only.

If the latter has a resistance R_1 it is easy to see that the V.A.F. will be given by the expression—

$$\text{V.A.F.} = \frac{\sqrt{R_1^2 + \omega^2 L_1^2}}{\sqrt{(R_1 + r_a)^2 + \omega^2 L_1^2}}\, mT.$$

In many practical cases the resistance of the primary R_1 may be neglected in comparison with $\omega^2 L_1^2$ and r_a.

The above case is one very commonly occurring in practice. Neglecting R_1, and keeping the frequency constant, it can be seen that the V.A.F. will approach the value mT as L_1 approaches infinity. When $\omega L_1 = r_a$, the V.A.F. will be given by

$$\text{V.A.F.} = \frac{mT}{\sqrt{2}} = 0{\cdot}707\, mT \doteqdot \frac{70}{100}\, mT.$$

For given values of L_1 and r_a, as the frequency increases ωL_1 becomes greater in comparison with r_a and the V.A.F. approaches the maximum value mT. As the frequency decreases the V.A.F. also drops, becoming about 70 per cent. of its value at high audio-frequencies, at a frequency obtained from $\omega L_1 = r_a$. Thus, if $r_a = 15{,}000\Omega$ and $L_1 = 50$ henries, we have

$$f = r_a/2\pi L_1 = 15{,}000/2\pi 50 = 49{\cdot}5 \text{ cycles.}$$

The value of L_1 is the value when there is no steady polarising current flowing in the winding. The presence of a steady current tends to decrease the value of L_1 (Vol. I), an effect which is countered by the adoption of "**parafeeding**" (*cf.* No. 49); this diverts the steady current through another path.

In the RADIO-FREQUENCY AMPLIFIER with transformer coupling, there is a considerable leakage because an iron core cannot, in general, be used, and hence it cannot be assumed that the voltage across the secondary is T times that across the primary.

(b) UNTUNED R/F TRANSFORMER COUPLING (Fig. 8).—If we assume, for purposes of simplification, that the resistance between grid and filament of the next valve is infinite, but take into account the inductance L_1 of the primary, the current variation I_a in the anode circuit is given by

$$I_a = \frac{mV_g}{\sqrt{r_a^2 + (\omega L_1)^2}}.$$

The voltage induced across the secondary circuit is $\omega M I_a$, and if we write $M = K\sqrt{L_1 L_2}$, where K is the coupling factor, the secondary voltage is given by

$$V'_g = \frac{\omega K\sqrt{L_1 L_2}\, mV_g}{\sqrt{r_a^2 + (\omega L_1)^2}}$$

and hence the V.A.F. $= \dfrac{m\omega K\sqrt{L_1 L_2}}{\sqrt{r_a^2 + (\omega L_1)^2}}$

If all the quantities with the exception of L_1 are kept constant (*i.e.*, L_2 is arbitrarily fixed), and this expression is differentiated with respect to L_1, the maximum value of the V.A.F. is found to occur when $\omega L_1 = r_a$. With this condition we have.

$$\text{V.A.F.} = \frac{mK}{\sqrt{2}}\sqrt{\frac{L_2}{L_1}}.$$

***(c)* R/F TRANSFORMER COUPLING, WITH A TUNED PRIMARY.**—From the above it would appear that the best value of V.A.F. is obtained when L_1 is small. If, however, L_1 is small, it becomes necessary to employ some artifice to increase the effective value of the impedance external to the valve. This is most conveniently done by adding a variable condenser in parallel with L_1 and arranging that this parallel circuit may always be tuned to resonance. We may then regard the effective resistance of the circuit L_1/C_1R_1 as being so much greater than the A.C. resistance of the valve that the whole of the voltage mV_g is applied across it. The oscillatory current flowing in L_1 is then given by $mV_g/\omega L_1$, and the voltage across the secondary is given by ωM times the value of this current.

$$\therefore \qquad V'_g = \frac{\omega M \,.\, mV_g}{\omega L_1} = \frac{mMV_g}{L_1} = \frac{mK\sqrt{L_1L_2}}{L_1}V_g.$$

As in the case of *(b)*, the value of R_g is assumed to be infinite.

Hence, the V.A.F. $= mK\sqrt{\dfrac{L_2}{L_1}}$

To give a large V.A.F., L_1 should be made small; there are, however, limits to this, because the smaller L_1 is made the smaller becomes the effective resistance L_1/C_1R_1, and hence r_a cannot be neglected by comparison, as was done above.

Alternatively, we might increase L_2, but, by so doing, its distributed self-capacity might be such as practically to short-circuit the grid-filament of the next valve, and reduce the V.A.F. very much. The effect of this self-capacity has been neglected in the above simplified discussion, but it is obviously of importance, especially at high frequencies.

As mentioned before, it is usually found best to make $L_2 = L_1$ in these radio-frequency transformer amplifiers, and avoid trying to increase amplification by transformer step-up effects.

***(d)* R/F TRANSFORMER COUPLING, WITH A TUNED SECONDARY** (*cf.* Fig. 12). By the simple coupled circuit theory of Vol. I, it has been shown that the total resistance and reactance in the primary of such a circuit is given by

$$R = R_1 + \frac{\omega^2 M^2 R_2}{Z_2^2} \qquad \text{and}$$

$$X = X_1 - \frac{\omega^2 M^2 X_2}{Z_2^2}$$

Since the secondary is tuned, we may put $Z_2 = R_2$ and $X_2 = 0$; moreover, we may neglect the effect of the inductance and resistance of the primary in comparison with r_a and the "reflected load" from the secondary.

Hence we write $Z_1 = \dfrac{\omega^2 M^2}{R_2}$. This gives

$$\text{Primary current} = I_1 = \frac{mV_g}{r_a + \dfrac{\omega^2 M^2}{R_2}}$$

$$\text{Induced secondary voltage} = \omega M I_1$$

$$\therefore \qquad \text{Secondary current} = I_2 = \frac{\omega M I_1}{R_2}$$

$$\text{Volts across secondary} = V'_g = \omega L_2 I_2$$

$$= \omega L_2 \,.\, \frac{\omega M}{R_2} \,.\, \frac{mV_g}{\left(r_a + \dfrac{\omega^2 M^2}{R_2}\right)}$$

$$\therefore \qquad \text{V.A.F.} = \frac{V'_g}{V_g} = \frac{\omega^2 m M L_2}{R_2 r_a + \omega^2 M^2}$$

If all quantities with the exception of M are kept constant, by differentiation it can be shown that the maximum value of V.A.F. will occur when $\omega^2M^2 = R_2r_a$; this defines the condition for "optimum coupling." With that condition we have—

$$\text{V.A.F.} = \frac{\omega^2 m M L_2}{2\omega^2 M^2} = \frac{mL_2}{2M}$$

$$\text{and } M = K\sqrt{L_1L_2} \quad \therefore \quad \text{V.A.F.} = \frac{mL_2}{2K\sqrt{L_1L_2}} = \frac{m}{2K}\sqrt{\frac{L_2}{L_1}}$$

This also gives the value of the coupling factor K; since $\omega^2 = 1/L_2C_2$, we have—

$$\omega^2 M^2 = R_2 r_a \quad \therefore \quad \frac{K^2 L_1 L_2}{L_2 C_2} = R_2 r_a \quad \ldots\ldots\ldots \quad \text{(optimum coupling)}$$

$$\therefore \qquad K^2 = \frac{R_2 r_a\, C}{L_1}$$

Taking practical values, this leads to values of K of the order of 0·8 ; this is considerably higher than the critical coupling " discussed in case (*e*) below ; clearly, it must be big since r_a is a large quantity.

Moreover, when $\omega^2M^2=R_2r_a$, the effective primary impedance becomes—

$$Z_1 = \frac{\omega^2M^2}{R_2} = r_a.$$

The last result demonstrates that **maximum V.A.F. is obtained when the effective valve output impedance is equal to the A.C. resistance of the valve.**

The remarks made in (*c*) about the value of L_2 apply also here. It should be noted that the simple treatment of (*b*), (*c*), and (*d*) shows that the amplification is dependent upon the frequency.

The forms of coupling in (*b*), (*c*) and (*d*) are particularly suitable for W/T receivers in which R/T reception is not a primary aim ; in the latter case, the desirability of " band pass " effects leads to the more complicated case outlined in (*e*) below.

***(e)* R/F TRANSFORMER COUPLING, WITH A TUNED PRIMARY AND SECONDARY** (Fig. 9).—This is a much more difficult case, and a full treatment cannot be included in this work.

Employing the usual symbols, if we take the case of an I/F band pass filter, the applied ω is fixed, $X_1=X_2 \doteqdot 0$, and M is the only variable quantity.

The equivalent circuit consists of an applied voltage mV_g, in series with the r_a of the valve and the primary tuned circuit, the latter being mutually coupled to the secondary. With the above assumptions, by analysis it can be shown that the secondary circulating current is given by—

$$I_2 = \frac{mV_g\left(\frac{M}{C_1r_a}\right)}{\sqrt{\left(R_1R_2 + \omega^2M^2 + \frac{r_a}{\omega^2C_1^2r_a}\right)^2 + \left(\frac{1}{\omega C_1r_a}\right)^2\left(R_1R_2 + \omega^2M^2\right)^2}} \qquad \ldots\ldots\ldots \quad (1)$$

If we make the further reasonable assumption that r_a is always greater than $1/\omega C_1$, the second term of the denominator may be regarded as a small quantity, and neglected. We then have

$$I_2 = \frac{mV_g\left(\frac{M}{C_1r_a}\right)}{R_1R_2 + \omega^2M^2 + \frac{r_a}{\omega^2C_1^2r_a}}$$

Now $\frac{1}{\omega^2C_1^2r_a} = \frac{L_1}{C_1r_a}$, since $X_1 = 0$ and $\omega^2 = 1/L_1C_1$.

Moreover, $\frac{1}{\omega^2C_1^2r_a}$ represents the **equivalent series resistance of the r_a of the valve transposed to the primary tuned circuit** (Vol. I, paragraph 310). Denoting this equivalent series resistance by R′ we may write

$$I_2 = \frac{(mV_g)\ \frac{M}{C_1r_a}}{(R_1 + R')\,R_2 + \omega^2M^2} \qquad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (2)$$

By differentiation it may be shown that I_2 is a maximum for variation of M when

$$\omega^2M^2 = (R_1 + R')\,R_2. \qquad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (3)$$

This defines the condition for **optimum coupling**, and it is interesting to observe that the problem has simplified down to that of two loosely coupled circuits tuned to the same frequency (Vol. I, paragraph 336), the primary being modified by having the equivalent series resistance of the r_a of the valve, placed in series with the resistance R_1 of the primary.

From this we may get the V.A.F. of the stage

Output volts across $C_2 = V'_g = \frac{I_2}{\omega C_2}$.

Substituting optimum values in (2).

$$V'_g = \frac{\frac{1}{\omega C_2}\left(\frac{\omega M}{\omega C_1r_a}\right)mV_g}{2\,(R_1 + R')\,R_2} \quad \ldots \quad \text{and if } C_1 = C_2$$

$$= \frac{\left(\frac{mV_g}{\omega^2C_1^2r_a}\right)\sqrt{(R_1 + R')\,R_2}}{2(R_1 + R')\,R_2} = \frac{mV_g\,R'}{2\sqrt{(R_1 + R')\,R_2}}$$

$$\therefore \quad \text{V.A.F.} = \frac{V'_g}{V_g} = \frac{mR'}{2\sqrt{(R_1 + R')\,R_2}} \qquad \ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (4)$$

Clearly, R′ cannot be taken as zero, for that would correspond with an infinite value of r_a, which would obviously give zero V.A.F. For any ordinary circuit R′ will be of the same order of magnitude as R_1, namely 5 ohms or so at about 600 kc/s. It is equally impossible to make R′ big, for that would be equivalent to shorting the tuned circuit with a low value of r_a.

From (3) we may obtain an expression for optimum coupling in more useful terms.

$$\omega^2 M^2 = (R_1 + R')R^2 = \left(R_1 + \frac{1}{\omega^2 C_1^2 r_a}\right) R_2$$

$$\therefore \quad K^2 = \frac{\omega^2 M^2}{\omega^2 L_1 L_2} = \left(\frac{R_1}{\omega L_1} + \frac{1}{\omega^3 L_1 C_1^2 r_a}\right)\frac{R_2}{\omega L_2}$$

and in terms of Q_1 and Q_2 (Vol I)

$$\left(\frac{1}{Q_1} + \frac{1}{\omega C_1 r_a}\right)\frac{1}{Q_2} = \left(\frac{1}{Q_1} + \frac{1}{Q_1'}\right)\frac{1}{Q_2}$$

where $Q'_1 = \frac{\omega L_1}{R'} = \frac{\omega L_1}{L_1/C_1 r_a}$

$$\therefore \quad K = \sqrt{\left(\frac{1}{Q_1} + \frac{1}{Q_1'}\right)\frac{1}{Q_2}} \quad \ldots\ldots\ldots\ldots\ldots \quad (5)$$

It will be noted that K is greater than $\frac{1}{\sqrt{Q_1 Q_2}}$, the value applicable to the case of two simple tuned coupled circuits, having the source of E.M.F. in the primary (Vol. 1). This is to be expected if one bears in mind the damping effect of the r_a of the valve; the correcting term $\frac{1}{Q'_1}$ must always be present.

It should be realised that formula (5), and others like it, only give an approximate idea of the requisite value of K; in practice, the optimum conditions are arrived at by a process of adjustment.

Good coils often have a Q of the order of 100; this gives the value of the critical coupling to be of the order of 0·01, or 1 per cent. In practice, it varies from 0·5 per cent. to somewhat over 1 per cent. Where good selectivity is essential, as in W/T receivers, the coupling is often made considerably less than the critical value. In this connection one may recall that **for couplings greater than critical coupling,** a double humped response curve results, and the two hump frequencies are given by

$$f_1 \doteq \frac{f_0}{\sqrt{1-K}} \quad \text{and} \quad f_2 \doteq \frac{f_0}{\sqrt{1+K}}$$

or

$$f_1 \doteq f_0\left(1 + \frac{K}{2}\right) \quad \text{and} \quad f_2 \doteq f_0\left(1 - \frac{K}{2}\right)$$

(by the binomial theorem, since K is small).

$$\therefore \quad f_1 - f_2 \doteq Kf_0 = \text{the peak separation.}$$

Hence, the peak separation approximately depends directly on the value of K.

In R/T work, where band pass effects are necessary (N.1, 31), intermediate frequency transformers are deliberately over-coupled beyond the optimum coupling.

***(f)* SOLUTION TO NUMERICAL PROBLEM No. 21** (End of Section).—The valve equivalent circuit [*Cf.* 19 (*a*), Fig. 10] should be used, the simple circuit being complicated by having the primary winding L_1 in parallel with the reflected resistance R_g/T^2. (*Cf.* Vol. I.)

$$\text{Primary current} = I_a = \frac{mV_g}{Z_{AB}}$$

$$\text{Transformer primary volts} = I_a Z_{XY}.$$

Hence, with unity coupling factor—

$$\text{Output volts} = V'_g = T\, I_a Z_{XY}$$

and substituting for I_a

$$\text{V.A.F.} = \frac{V'_g}{V_g} = \left(\frac{Z_{XY}}{Z_{AB}}\right) mT \quad \ldots\ldots\ldots\ldots\ldots \quad (1)$$

To find Z_{AB} it is necessary to determine the reactive and resistive components of the impedance Z_{XY}, a matter most conveniently done by "j" methods (Vol. I).

Let $R_g/T^2 = R_1$.

$$\frac{1}{Z_{XY}} = \frac{1}{j\omega L_1} + \frac{1}{R_1} = \frac{\omega L_1 - jR_1}{\omega L_1 R_1} \quad \ldots\ldots\ldots\ldots\ldots \quad (2)$$

$$\text{Hence, } Z_{XY} = \frac{\omega L_1 R_1(\omega L_1 + jR_1)}{\omega^2 L_1^2 + R_1^2} = \underset{\text{(Resistive)}}{\frac{\omega^2 L_1^2 R_1}{\omega^2 L_1^2 + R_1^2}} + \underset{\text{(Reactive)}}{\frac{j\omega L_1 R_1^2}{\omega^2 L_1^2 + R_1^2}} \quad \ldots\ldots\ldots\ldots\ldots \quad (3)$$

Also from (2) we get the modulus of Z_{XY}

$$= |Z_{XY}| = \frac{\omega L_1 R_1}{\sqrt{R_1^2 + \omega^2 L_1^2}} \quad \ldots\ldots\ldots\ldots\ldots \quad (4)$$

At 100 cycles:

$$R_1 = 55{,}555 \text{ ohms}; \quad \omega L_1 = 18\cdot 84 \times 10^3 \text{ ohms.}$$
$$R_1^2 = 31\cdot 1 \times 10^8 \text{ ohms}; \quad \omega^2 L_1^2 = 3\cdot 55 \times 10^8 \text{ ohms.}$$

Substituting in (3)

$$Z_{XY} = \frac{3\cdot 55 \times 10^8 \times 55{,}555}{(3\cdot 55 + 31\cdot 1) \times 10^8} + \frac{j31\cdot 1 \times 18\cdot 84 \times 10^{11}}{(3\cdot 55 + 31\cdot 1)\, 10^8}$$
$$= 5{,}680 + j16\cdot 92 \times 10^3. \quad \ldots\ldots\ldots\ldots \quad (5)$$

$$\therefore \quad |Z_{XY}| = \sqrt{5{,}680^2 + 16\cdot 92^2 \times 10^6} = 17\cdot 82 \times 10^3 \text{ ohms.}$$

Now from (5)

$$Z_{AB} = (5{,}680 + 30{,}000) + j16\cdot 92 \times 10^3$$
$$= 35{,}680 + j16\cdot 92 \times 10^3,$$

$$\therefore \quad |Z_{AB}| = \sqrt{(35\cdot 68^2 + 16\cdot 92^2) \times 10^6} = 39\cdot 5 \times 10^3 \text{ ohms.}$$

From (1)

$$\text{V.A.F.} = \frac{17\cdot 82 \times 10^3}{39\cdot 5 \times 10^3} \times 120 = 54\cdot 2$$

At 10,000 cycles:—

Here $\omega^2 L_1^2 = 35{,}500 \times 10^8$, and is large in comparison with R_1^2; the latter may be neglected by comparison. This makes Z_{XY} into a pure resistance. From (4)

$$Z_{XY} = R_1 = 55{,}555 \text{ ohms.}$$

Hence,

$$Z_{AB} = 85{,}555 \text{ ohms.}$$

$$\therefore \quad \text{V.A.F.} = \frac{55{,}555}{85{,}555} \times 120 = 77\cdot 9.$$

The simple formula in *(a)* leads to the same result.

20. **Permeability Tuning.**—Up to about 1933, the tuning of circuits was carried out principally by means of variable condensers and fixed inductances; this provided the most convenient method leading to one knob tuning controls, using assemblies of ganged condensers. In spite of its apparent advantages, the system had certain drawbacks, but **variable inductance tuning** was not seriously considered, as an alternative, until the introduction of the low loss high permeability powdered iron cores made it possible to vary the inductance of a coil by altering the relative positions of coil and core. The inductance of such a coil has its smallest value when the core is completely removed, and has a maximum value when the reluctance of the magnetic circuit is reduced by inserting the core to its full amount. The system has become known as **permeability tuning,** although the inductance is controlled by altering the *reluctance* of an air gap, in order to produce corresponding alterations of flux.

Before the invention of these special cores, the most commonly used continuously variable inductance was the variometer. The latter suffered from the disadvantage that the ratio of inductance to H/F resistance varied greatly over the tuning range; it can be shown that permeability tuning represents an approach to the ideal system, provided that the ratio L/R is kept constant. Under the best conditions it leads to an increase of **selectivity** with frequency, since $Q = \omega L/R$; but the value of Q affects the **sensitivity** of the receiver, which for some purposes should be independent of the frequency.

The construction of the cores is a technical matter, the details of which might vary, and will not be fully described here. In many cases, the core consists of particles of iron, often colloidal in size and nature, embedded in an insulating material by the application of high pressure. In effect, any process produces a core with the properties of iron, but in which eddy currents cannot flow because of the insulating surfaces separating the particles; hysteresis and eddy current losses are thereby minimised.

The defects of ordinary condenser tuning become clearer if one considers what happens in an R/F amplifying stage employing a tuned anode coupling to the next valve. The **stage gain** depends upon the value of L/CR. Over any one tuning range, if L and R are constant, the value of L/CR is inversely proportional to the capacity C. This produces alterations in the amplification which may, in some measure, be offset by corresponding changes in R; as C is *decreased*, to increase the frequency, the value of R usuallv *increases* and might annul the alteration in C.

Moreover, the **selectivity** of a tuned circuit is proportional to the coil amplification Q, and is given by

$$Q = \frac{\omega L}{R} = \frac{1}{R}\sqrt{\frac{L}{C}} = \frac{1}{\sqrt{R}}\sqrt{\frac{L}{CR}}$$

If L/CR is maintained at a fixed value, in order to achieve **constant sensitivity,** it is clear that, with ordinary condenser tuning, the value of R increases and the selectivity must decrease with increasing frequency. Alternatively, the selectivity may be maintained constant, but in that case the sensitivity will vary.

If inductance tuning is employed, R decreases when L is decreased in order to raise the frequency, since **the ratio L/R is constant** ; this implies that the **selectivity must increase with increase of frequency,** an essential advantage which is claimed for permeability tuning in receivers mainly employed in R/T reception.

The formula for selectivity gives results on a percentage basis of mis-tuning, two receivers working on different frequencies being equally selective if they give equal attenuation at a certain percentage frequency away from resonance. For R/T work, with side bands covering a frequency band (say) 8 kc/s. in width on each side of the carrier (N.15-18), it is advantageous that the selectivity should be the same at a fixed number of kc/s. from resonance, whatever the resonant frequency may be. This desirable result may be achieved if the selectivity of the receiver increases with increasing frequency, as it is claimed that it does with permeability tuning.

21. **A/F Amplification. The Note Magnifier.**—It has already been observed that the essential principles of A/F amplifiers are the same as those of H/F amplifiers. Examples of each of the different types of coupling will be seen ; for high fidelity work there will be an evident preference for resistance-capacity coupling, but for large power output the transformer coupling remains the most popular, since it is the only one in which the V A.F. of a valve stage may be greater than *m*.

In early Service receivers, note magnifiers were generally separate instruments, but in modern Service receivers, as in all commercial ones, the requisite A/F and R/F amplification stages are generally incorporated in the same model.

In cases where the Service note magnifier is a separate instrument, the first stage inside the box is intended to be the inter-valve coupling between the detector valve and the first valve of the A/F amplifier ; as such, this coupling should be connected between the anode of the detecting valve and the positive of the H.T. battery. For testing purposes, telephones could be inserted at the detector stage by connecting a telephone transformer and condenser instead of completing the connections to the note magnifier.

Service note magnifiers designed to share common batteries with other amplifiers have only one input terminal ; this is connected to the anode of the detecting valve, and the circuit is completed through the common anode battery.

Examples of note magnifiers will be seen in later drawings showing complete receivers.

22. **Power Amplification.**—So far we have been concerned with output circuits designed to give as high a V.A.F. as possible, so that the maximum voltage may be applied to the next valve in the amplifier.

In the case of the last valve of a note magnifier, however, in the anode circuit of which is connected as output circuit the telephones or loudspeaker, it is important to obtain maximum power output. This implies large current variations as well as large voltage variations, and the best condition is when the product of the oscillatory current in the anode circuit and the oscillatory voltage set up across the output circuit is a maximum. For a triode valve, it is shown in the following paragraphs that this is the case when the impedance in the external circuit is equal to the A.C. resistance (r_a) of the valve, provided *maximum* power dissipation alone is considered ; if distortion of the power output is to be avoided, the external impedance should be about twice the A.C. resistance of the valve (*see below*).

Valves used for power have, generally, A.C. resistances of from 1,000 to 5,000 ohms, and are designed to be used with higher anode voltages than usual. Their amplification factor is small

(from 2 to 10), and their mutual characteristics are therefore straight over a large range of negative grid voltage, so that considerable negative grid bias may be applied and distortionless amplification be secured for large values of grid-filament voltage variation.

★23. **Maximum Power Output for a Triode.**—Let us suppose that the resistance of the output stage is given by R.

As before, $I_a = \dfrac{mV_g}{r_a + R}$, and the output voltage across R is $\dfrac{mR\,V_g}{r_a + R}$.

The product gives the output power $= \dfrac{Rm^2V_g^2}{(r_a + R)^2}$. We want R to be such that this is a maximum.

Hence, differentiating $\dfrac{R}{(R + r_a)^2}$ with respect to R, and equating to zero, we obtain :—

$$(R + r_a)^2 - 2R\,(R + r_a) = 0$$

or

$$R + r_a = 2R,$$

$$\therefore \quad R = r_a \qquad (Cf.\ R.35,\ N.56).$$

Therefore, for maximum power output, the external resistance should equal the impedance of the valve.

The impedance of a telephone receiver or loudspeaker cannot, of course, be regarded as being purely a resistance, but the above result leads to the practical case, in which it can be arranged that the A.C. resistance of the last valve is about equal to the impedance of the receiving instrument.

★24. **Maximum Power with Distortionless Amplification.**—In the above argument, no account is taken of the fact that when the grid wing extends to the limits of the dynamic characteristic, grid current will flow during part of the positive half-cycle of grid voltage and the wave form of the output will be distorted. For distortionless amplification, the optimum value of the external impedance is about twice the A.C. resistance of the valve, as will now be proved.

Fig. 11 represents the working conditions. V_0 is the H.T. voltage on the anode under static conditions, and the corresponding static characteristic is shown. The curve labelled $V_a = V_1$ is the static characteristic for which flow of grid current commences at the same grid voltage as lower-bend curvature. V_1 is found, of course, by inspection of the family of mutual characteristics.

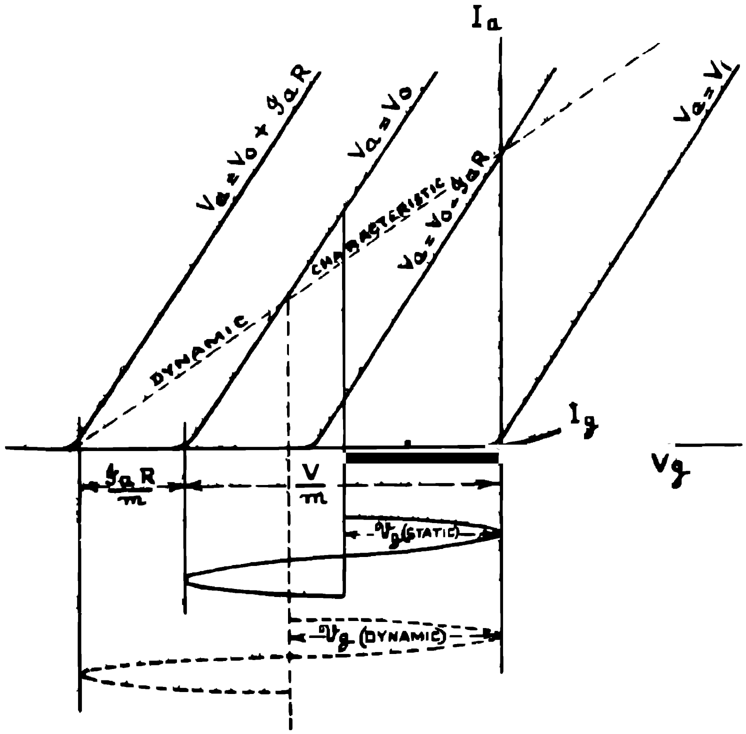


FIG. 11.

If $V_0 - V_1 = V$, the corresponding range of grid voltage over which no distortion occurs under static conditions is $\dfrac{V}{m}$, where m is the amplification factor of the valve (*Cf.* B.34, B.38, Fig. 24).

If the external impedance has an effective resistance R, then, with an oscillatory current of amplitude $\mathscr{I}_a$, the variation of anode voltage is from $V_0 - R\mathscr{I}_a$ to $V_0 + R\mathscr{I}_a$. The static characteristics corresponding to these voltages are shown, and also the dynamic characteristic on which the valve is operating. It can easily be seen that, compared with static conditions, the range of grid voltage for distortionless working has been increased by an amount equal to $\dfrac{R\mathscr{I}_a}{m}$, provided that the working point is moved to the left by increasing the negative bias the appropriate amount.

The total permissible grid swing is therefore $\dfrac{V + R\mathcal{I}_a}{m}$. The optimum grid bias and the permissible peak grid voltage are both half of this, *i.e.*, $\dfrac{V + R\mathcal{I}_a}{2m} = \mathcal{V}_g$.

The slope of the dynamic characteristic is $\dfrac{m}{R + r_a}$, ... (paragraph 4),

$$\therefore \mathcal{I}_a = \frac{m}{R + r_a}\mathcal{V}_g = \frac{m\,(V + R\mathcal{I}_a)}{2m\,(R + r_a)} = \frac{V + R\mathcal{I}_a}{2(R + r_a)},$$

$\therefore \mathcal{I}_a = \dfrac{V}{2r_a + R}$. ...... We have $I_a = \dfrac{\mathcal{I}_a}{\sqrt{2}}$, and hence the power (P) dissipated in the external impedance is given by

$$P = RI_a^2 = \frac{R\mathcal{I}_a^2}{2} = \frac{RV^2}{2\,(2r_a + R)^2}.$$

As in the previous paragraph, the value of R for maximum power is obtained by differentiating this expression and equating the result to zero. This gives

$$(2r_a + R)^2 - 2R\,(2r_a + R) = 0,$$
$$\therefore 2r_a + R - 2R = 0,$$
$$\therefore R = 2r_a.$$

The expression for the best negative grid bias in terms of known quantities, and assuming that grid current starts to flow at zero grid volts, is given by

$$\frac{V + R\mathcal{I}_a}{2m} = \frac{V + \dfrac{RV}{2r_a + R}}{2m} = \frac{(r_a + R)\,V}{m\,(2r_a + R)}.$$

When $R = 2r_a$ this becomes $\dfrac{3V}{4m}$.

The modification required if grid current starts to flow at some other grid voltage is obvious.

25. **A Complete Receiver ; Tuner, R/F Amplifier, Note Magnifier.**—Fig. 12 is an equivalent diagram of a receiver designed for use on the L/F range ; it consists of a tuner followed by three stages of R/F amplification, a detector valve, and three stages of note magnification. It is a simple **T.R.F.** (tuned radio frequency) **type receiver.**

The tuner has two positions, " Stand-by " and " Tune." In the Stand-by position, which is used when searching for signals, the aerial tuning inductance is directly across the grid and filament of the first valve, and the secondary circuit L_2 C_7 is cut out. This position is therefore unselective. When the required signal has been found, the switch is made to the Tune position ; this brings the secondary circuit into use and increases the selectivity.

The central stud of the Tune—Stand-by switch is earthed to prevent the switch from acting as a capacitive coupling between aerial and secondary circuits.

It will be seen that, when the tuner is in the " Stand-by " position, the grid-filament capacity of the first valve is in parallel with L_1. In the " Tune " position, this capacity is in parallel with the secondary tuning condenser C_7. To preserve the aerial tuning, it is therefore necessary to insert, in parallel with L_1, a condenser whose capacity is the same as the grid-filament capacity of the valve. This is called the Valve Equivalent Condenser (V.E.C.). A_1 and A_2 are lightning arresters to protect the receiver from damage. The aerial condenser has two positions, series and parallel, according to the frequency which is being received.

V_1, V_2 and V_3 act as radio-frequency amplifiers, V_4 as the detector.

The inter-valve coupling is by means of the transformers T_1, T_2, T_3, whose secondaries are tuned by the condensers C_1, C_2 and C_3.

The grid potentials of the second and third valves can be made more or less positive by means of the potentiometer R_3, which is connected across the filament battery terminals. So far, the necessity of having the grid negative for distortionless amplification and efficiency has been emphasised, but we shall see, in dealing with self-oscillation, that the grid may be given positive bias, with consequent loss of efficiency, as a means of preventing self-oscillation.

The rheostat R_2 controls the filament current of all the valves in parallel.

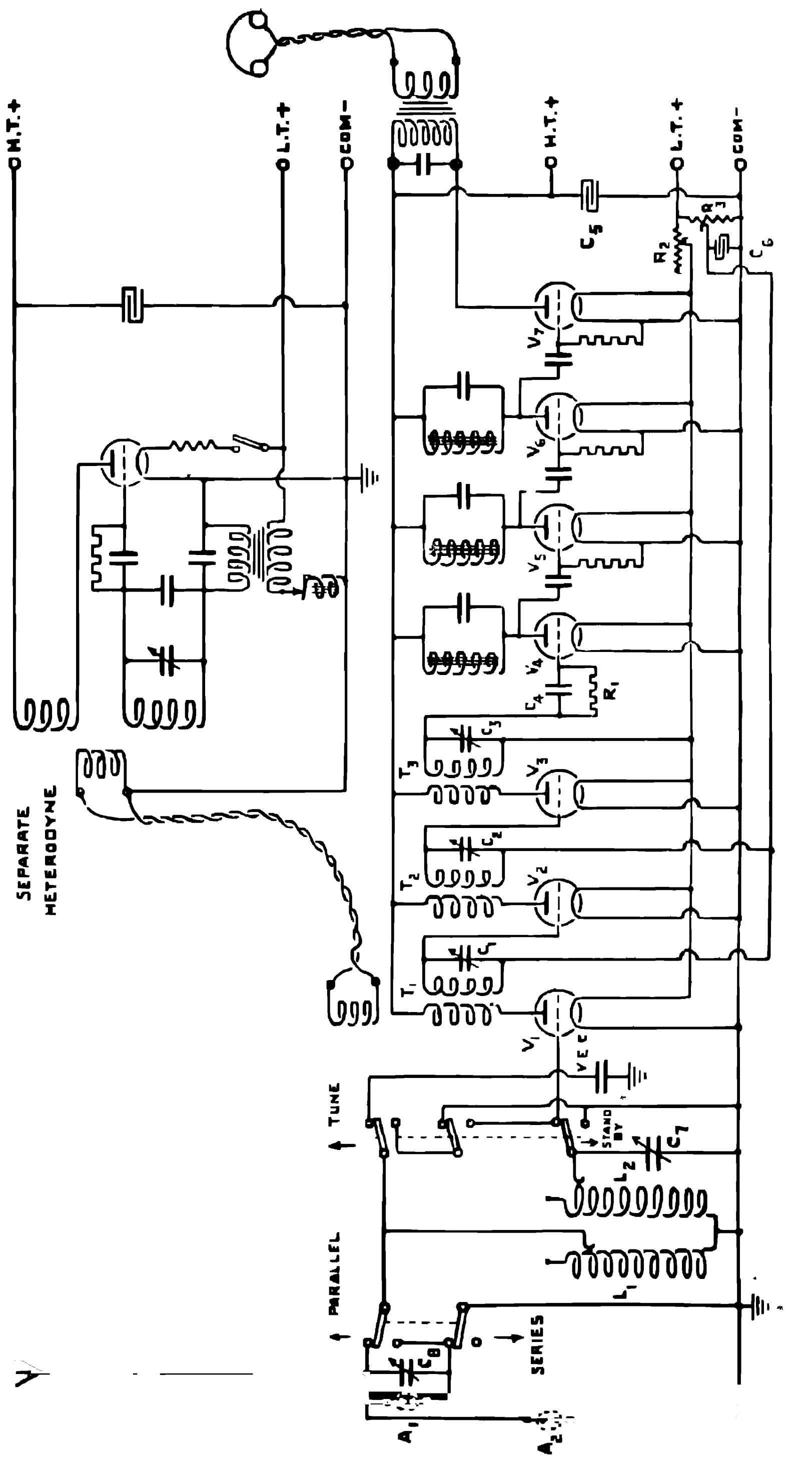

H.T.+
L.T.+
COM-
SEPARATE HETERODYNE
TUNE
PARALLEL
SERIES
STAND BY
V_1
V_2
V_3
V_4
V_5
V_6
V_7
T_1
T_2
T_3
C_1
C_2
C_3
C_4
C_5
C_6
C_7
C_8
R_1
R_2
R_3
L_1
L_2
A_1
A_2

FIG. 12.

In order to provide for cumulative grid detection, the condenser C_4 and leak R_1 are connected to the grid of valve V_4, which is given positive bias by connecting its input circuit to the positive L.T. lead.

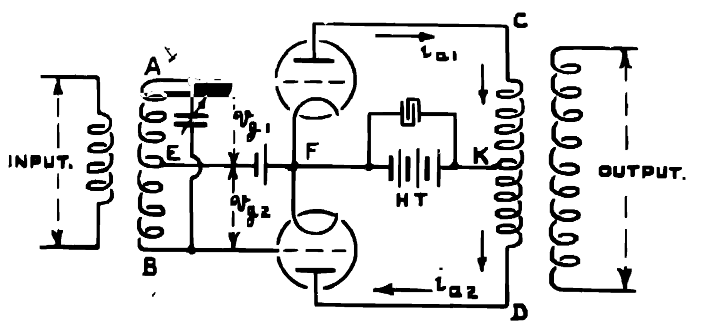
The condensers C_5 and C_6 (capacity 2μ F) are put in shunt with the amplifier batteries and the grid potentiometer respectively, to carry both radio and audio-frequency variations, and prevent self-oscillation, which might arise as a result of coupling between different parts of the amplifier through the resistances of the batteries.

The three stages of note magnification V_5, V_6, V_7, utilise iron cored tuned choke capacity coupling. In the output circuit of the power stage is the telephone transformer and its associated circuit.

26. **Valves in Push-Pull.**—In addition to the series and parallel arrangements of valves there are many circuits to which the name "push-pull" has been applied, although in certain cases the name "push-pull" might be more appropriate.

Push-pull circuits are finding an increasing number of applications, but it is here proposed only to consider their applications to A/F power amplification. They are, however, extensively used as R/F amplifiers (N.66, K.34, 41, 45, T.10), and as heterodyne detectors (T.10). Other push-pull applications will be discussed as they arise.

27. **Class "A" Push-Pull A/F Amplifiers.**—A/F amplifiers only differ in small details from R/F amplifiers, a principal one being that the former employ iron-cored transformers. Fig. 13 represents a basic form of the circuit in which, for simplicity, the iron cores have been omitted. The input circuit is, of course, the output circuit of the previous stage of amplification, and is shown as tuned-transformer coupled. The secondary coil AB, instead of being across grid and filament of one valve, as in cascade amplification, is connected across the grids of two valves, as shown. The electrical mid-point E of AB is common, via the grid bias battery, with the filaments of the two valves. The output circuit illustrated is of the same type as the input and is connected across the anodes of the two valves. The H.T. battery is inserted between the electrical mid-point of the output primary and the common filament connection. It will thus be seen that the arrangement is a very symmetrical one, if the valves are matched so as to have the same characteristic curves. It is this essential symmetry which tends to treat each half cycle of the input signal in the same manner, thereby producing strong **attenuation of all even harmonics.** This is especially useful with triodes since the second harmonic is usually particularly evident (*cf.* paragraph 29).



PUSH-PULL ARRANGEMENT OF VALVES

FIG. 13.

An oscillatory voltage is produced across AB by the action of the incoming signal on the preceding stages of amplification. Thus with respect to E, whenever A is at a certain positive potential, B is at an equal negative potential and *vice versa*; *i.e.*, the oscillatory P.Ds. from A to E and from B to E are always equal and opposite. These are the grid-filament P.Ds. applied to the two valves. Hence V_{g1} and V_{g2} are equal in amplitude, but 180° out of phase with each other.

Now consider the output circuit. When no signal is being received, a steady electron current is flowing from filament to anode of each valve, its value depending on the steady anode and grid voltages. The currents in the two valves are obviously equal, if the valves are perfectly matched and the H.T. battery positive tap is at the electrical mid-point of the output circuit. As regards the output coil, however, these two equal currents flow in opposite directions through the two halves

of the coil. The steady fluxes they produce in the coil are thus equal and opposite, and there is no resultant flux, and no consequent magnetisation of the core in the case of A/F amplifiers. The advantages of this in preventing distortion of the wave form will be evident from Vol. I.

When a signal is being received, equal grid voltages in anti-phase are applied to the two valves. If i_{a1} and i_{a2} are the oscillatory anode currents produced in consequence, these two currents will also be equal and opposite, provided that the grid swings do not extend beyond the straight portions of the dynamic characteristics of the valves, and that grid current is not allowed to flow (Fig. 14 (*a*)). The resulting oscillatory P.Ds. produced across the two halves of the output coil from C to K and D to K are therefore also equal and opposite (180° out of phase), and so the P.Ds. produced in the coil

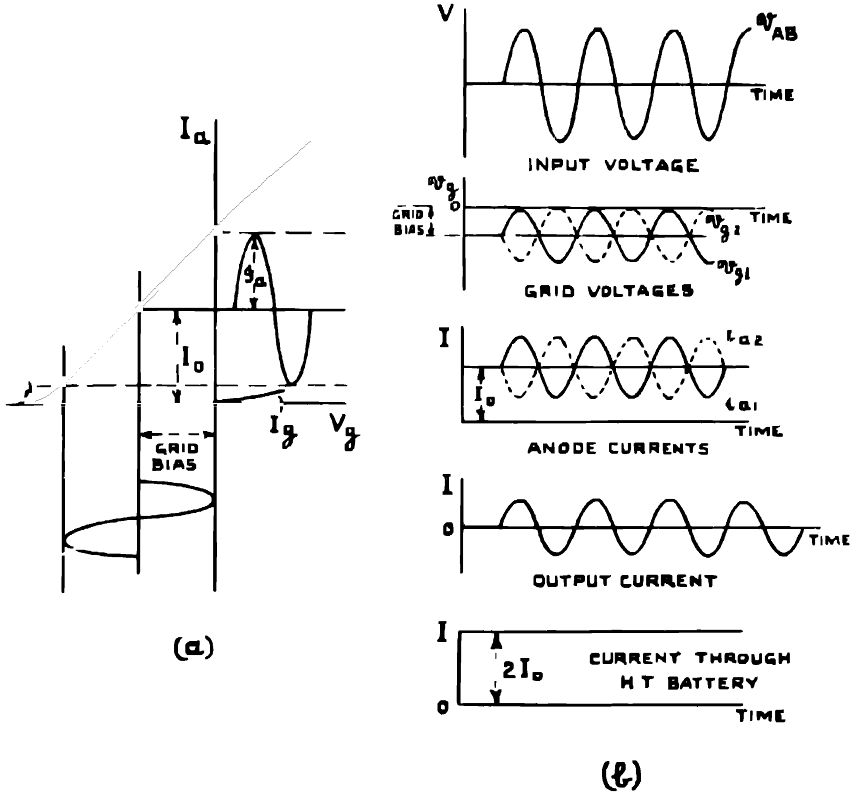


FIG. 14.

from C to K and K to D are equal and **in the same direction (in phase)**. In other words, the oscillatory current in the two halves of the coil is the same both as regards amplitude and phase, which is **equivalent to one oscillatory current of the same value throughout the transformer winding** (Fig. 14 (*b*)). In paragraph 29 it is shown that the fundamental components are additive, and the second harmonics cancel each other. In general, **the circuit attenuates all even harmonics**.

On the other hand, i_{a1} and i_{a2}, flowing through the H.T. battery and common lead, are additive, and, being equal and opposite, cancel each other out, *i.e.*, no resultant oscillatory current flown in this lead, and the possibility of back coupling from this stage to earlier stages through the common H.T. battery and anode lead is avoided, with beneficial results on the stability of the amplifier.

In the case discussed above, the best grid bias for distortionless amplification is the voltage half-way between the lower bend of the dynamic characteristic and the voltage at which grid current starts to flow (Class " A " conditions, or **mid-point biasing**).

As only half the total input voltage is applied between grid and filament of each valve, the push-pull arrangement will give distortionless amplification of twice the oscillatory input voltage

that either valve would deal with singly as a stage in a cascade amplifier, and so is particularly useful in the later states of A/F amplification where the input voltage is large.

28. **Class " B " Push-Pull A/F Amplifier. Curvature Biasing** (Q.P.P.).—The input voltage of a push-pull arrangement can be approximately doubled without distorting the output by increasing the grid bias until each valve is operating on **the mid point of the bottom bend** of its mutual characteristic. Under these conditions, with no input signal, the anode current is very small and produces a corresponding economy in the use of battery power ; one of the chief advantages of the system is this **economy in battery power,** an essential feature in portable receivers.

A class " B " amplifier is one which is operated with such a grid bias that there is practically no D.C. anode current when the signal voltage is zero (*cf.* paragraph 8, and N.45, 49). This is a general definition of Class " B."

With this arrangement, each valve *singly* would act as a rectifier, for during the negative half cycle of grid voltage practically no current will flow. In the *push-pull* circuit, however, the negative half cycle on one grid is the positive half cycle on the other, since the grid voltages are in anti-phase. The result is that during one complete cycle of input voltage, one valve amplifies the positive half cycle without distortion, and the other valve does the same for the negative half cycle. Each valve is quiescent, or out of action, for one half cycle, and for this reason the name " **quiescent push-pull** " (Q.P.P.) has been applied to the arrangement ; it may be noted that the circuit illustrates the idea of two valves working in *push-push*, each valve pushes in turn, and is essentially different from the action of the Class " A " circuit. The process is illustrated in Fig. 15.

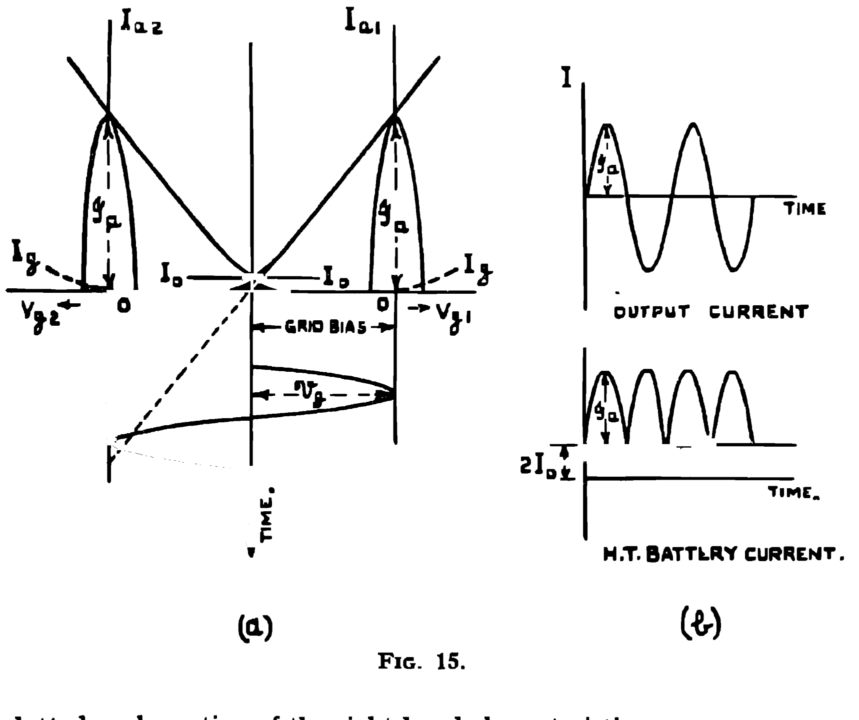


FIG. 15.

To prevent distortion the two valves must have identical characteristics. In the diagram, the two characteristics are shown in opposite directions to take account of the fact that positive oscillatory grid voltages on one valve correspond to negative ones on the other. It will be seen from the figure that the curvatures at the bottom of each characteristic cancel each other out as regards incoming signals, and the nett effect is to give a straight characteristic with zero flow of grid current over more than twice the range of either valve independently. This is indicated by the dotted prolongation of the right-hand characteristic.

This was the original method employed in push-pull circuits, but it suffers from the following disadvantages compared with the " mid-point biasing " scheme :—

(*a*) During the negative half cycle for either valve, its A.C. resistance is extremely large. The positive half wave of oscillatory anode current, produced in one valve thus flows through half the external circuit, and then returns to filament via the H.T. battery. During the negative half cycle the same process occurs with the other valve, Fig. 15 (*b*). Thus, although the steady current from the H.T. battery is

smaller with this method, the oscillatory current through it is large. The advantages possessed by the Class " A " push-pull arrangement for preventing back coupling through the H.T. battery are therefore sacrificed.

(*b*) The grid bias is critical, and demands a careful study of the characteristics to find the correct operating points. In addition, the valves must be much more carefully matched than in the mid-point biasing system, particularly as regards the bottom bends of their characteristics, and this is the point where deviations from uniformity in the same type of valve are most pronounced. Further, with mid-point biasing, a difference in the slope of the mutual characteristics of the two valves will not produce distortion. The alternating flux-linkage with the output coil is proportional to the product of the number of turns and the alternating current. Though these are different for the two halves of the coil, each half still produces flux-linkage whose wave form is a faithful reproduction of the input voltage wave over a complete cycle, and the only effect is that the amplitude of resulting flux-linkage is the mean of the flux-linkages produced in the two halves of the coil, and not exactly equal to either. Different slopes, however, will mean that the oscillatory currents flowing through the battery from the two halves of the coil are not equal in amplitude, so that there will be a resultant oscillatory current through the battery, with possibilities of back coupling. Hence only slight differences in characteristic slope should be permitted, and the H.T. battery should always be by-passed by a condenser.

With bottom bend biasing it is easily seen that different slopes of the characteristics will mean that the amplitude of the positive half-cycle of flux-linkage is different from the amplitude of the negative half-cycle, and so distortion is produced.

(*c*) The grid swing is doubled and gives approximately twice the anode current that would be produced from each valve working under Class " A " conditions. In each half cycle the anode current only flows in one half of the primary output winding; if the impedance matching arrangements fully load the stage to give maximum undistorted power output, it is easy to see that the power developed by two valves worked under Class " B " conditions (Q.P.P.) may approach twice the output of the same valves arranged for Class " A " operation (paragraph 31).

★29. **Mathematical Note on Push-Pull Amplifiers.**—The input E.M.F. is applied by means of a transformer having a centre tap connected to the filament; this gives rise to two E.M.F.'s opposite in phase, and equal in magnitude to half of the total input E.M.F. The latter are applied between grid and filament of the two valves, and we may write—

$$V_{g1} = A \sin \omega t \text{ and } V_{g2} = -A \sin \omega t.$$

Assuming that the valve characteristics are not quite linear, the anode currents and grid potentials may be related by an equation of the form $i_a = aV + bV^2$, where a and b are constants. Then

$$\begin{aligned} i_{a1} &= aA \sin \omega t + bA^2 \sin^2 \omega t \\ &= aA \sin \omega t + \tfrac{1}{2}b\, A^2 - \tfrac{1}{2}b\, A^2 \cos 2\omega t \\ &\quad \text{(Fundamental)} \qquad\qquad \text{(2nd harmonic)} \end{aligned}$$

and
$$i_{a2} = -aA \sin \omega t + \tfrac{1}{2}b\, A^2 - \tfrac{1}{2}b\, A^2 \cos 2\omega t.$$

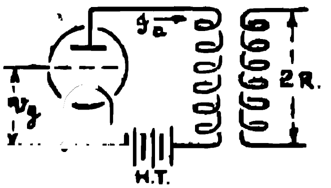
Since the output primary winding is arranged so that the fundamental oscillatory currents produce effects which *assist* each other, it is evident that the two **second harmonic terms** produce flux with the secondary which **annul each other.**

With more curvature in the characteristic higher harmonics will be produced (*cf.* N.18), but in all cases this coupling leads to **cancellation of the even harmonics** in the secondary coil of the output transformer. This is one of the chief advantages of the push-pull circuit, since, with triodes, the second harmonic is usually greater than the higher ones.

It is also evident that in the H.T. battery lead the fundamental components cancel out but the even harmonics are additive (*cf.* paragraph 28).

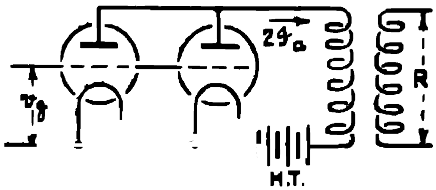
30. **Valves in Parallel. Impedance Matching.**—The advantages for power amplification of a push-pull arrangement may be seen by comparing it with the use of valves in parallel, an alternative method which may be used when it is necessary to obtain more power output than one valve can give.

The system is shown in Fig. 17 which should be compared with Fig. 16, and the push-pull arrangement shown in Fig. 13. To facilitate comparison, the output circuits are drawn as shown, and not in accordance with the standard method used in other figures.



SINGLE POWER VALVE

FIG. 16.



TWO POWER VALVES IN PARALLEL

FIG. 17.

The single power valve is one in which we may assume the amplitude of the permissible grid swing is $\mathcal{V}_g$, giving an R.M.S. anode current of I_a.

The output circuit is arranged in each case to give maximum undistorted power under these conditions. The internal A.C. resistance of the two valves in parallel is half the A.C. resistance of either valve working singly, and so the effective resistance in the anode circuit must be half that in the single power valve circuit. For this reason the output impedances are labelled R and 2R respectively in Figs. 16 and 17.

The two valves in parallel can only be allowed the same amplitude of grid swing ($\mathcal{V}_g$) as the single power valve ; the total anode current is, however, doubled, and this gives double the power output which is obtainable in the single power valve case (paragraph 31).

In the Class " A " push-pull arrangement, the A.C. circuit is through the two valves in series and the external anode circuit. The total internal A.C. resistance is thus twice that of either valve alone, and so the external effective resistance should be four times that used for the parallel combination, or twice that of the single power valve, *i.e.*, 4R.

This push-pull arrangement can cope without distortion with twice the permissible total input voltage ($2\mathcal{V}_g$) of the single power valve case, the input to each valve being $\mathcal{V}_g$.

Under the same operating conditions, the power output of two valves in Class " A " push-pull is the same as with the same two valves in parallel (paragraph 31). Second harmonic distortion is, however, less in the former case, so we may say, that for a given percentage of second harmonic the push-pull amplifier gives relatively more power than can be obtained by using the same two valves in parallel.

31. **Comparison of Various Forms of Power Output Stage.**—It is clearer to give a symbolical form to the above comparisons. In each case we may consider that the dynamic characteristics, like the static ones, have the same slope g'_m, an oscillatory input of R.M.S. value V_g, giving an R.M.S. oscillatory anode current of I_a in each valve ($= g'_m V_g$). In each case it is assumed that the stage is fully loaded for maximum undistorted output, using the same matching transformer.

(*a*) SINGLE POWER VALVE.—The input voltage (R.M.S.) is V_g, and so the anode current is I_a.

Hence the power obtained is proportional to $2R \times (I)_a^2 = 2R\, I_a^2$.

(*b*) TWO VALVES IN PARALLEL.—The input voltage is V_g. The anode current is I_a from each valve, *i.e.*, a current of R.M.S. value $2I_a$ flows in the external circuit.

Hence the power obtained is proportional to $R \times (2I_a)^2 = 4R\, I_a^2$.

(c) TWO VALVES IN CLASS " A " PUSH-PULL.—The input voltage to each valve is V_g, the total input voltage being $2V_g$. As already seen (paragraph 27), the current in the external circuit is I_a, produced respectively in the two halves of the transformer winding by the two valves ; the relative anode to anode load is now 4R.

Hence the power obtained is proportional to $4R \times (I_a)^2 = 4R\ I_a^2$.

(d) TWO VALVES IN CLASS " B " (Q.P.P.) PUSH-PULL.—The input voltage to each valve is now $2V_g$, the total input volts being $4V_g$ In each half cycle the current $2I_a$ only flows in one half of the output winding, and works into an effective anode to filament impedance proportional to 2R.

Hence the power obtained is proportional to $2R\ (2I_a)^2 = 8R\ I_a^2$.

This power is developed by each valve in alternate half cycles of grid input voltage, from which it appears that the power output of two valves in Class " B " (Q.P.P.) may approach twice the output of the same valves under Class " A " conditions. It is shown, later, that the *efficiency* with which this power conversion is achieved is considerably greater than in the Class " A " case (K.12, etc.).

Since each valve works singly into the whole load, in effect only one half of the primary winding of the matching transformer is being used during each half cycle. In comparison with the preceding cases, the effective value of the transformer ratio is therefore doubled A relative anode to anode load of 8R will then be equivalent to $8R/(2T)^2$, giving an effective load on each valve proportional 2R.

(e) TWO VALVES IN CLASS " A-B " (LOW LOADING) PUSH-PULL.—This represents an operating condition midway between Class " A " and Class " B," the anode to anode load being less than that employed in normal Class " A " amplification, in order to increase the power output at the expense of some additional harmonic distortion (N.48).

When fully loaded, each valve singly works for slightly more than one-half cycle. The grid bias is adjusted so that the valves draw more anode current than in the Class " B " condition, but less than when adjusted for Class " A."

Naturally, the efficiency and relative power output of the Class " A-B " amplifier is somewhere between that obtainable in the simple Class " A " and Class " B " cases. One of its chief advantages is that its actual operation appears to vary between the pure Class " A " and Class " B " states as the input varies from small values to high values ; one of the major objections to Class " B " amplification consists in the distortion introduced with low signal inputs, a matter partially remedied in this way.

The use of Class " A-B " is becoming increasingly popular where large power output is required, and a little distortion is acceptable.

Power amplification using two valves in parallel, or one single valve, suffers from the following serious disadvantages :—

(a) A large steady current flows in the output primary coil, or choke. With a winding similar to that which could be used in the push-pull arrangement, the core would probably be saturated. In any case, the effective inductance of the coil would be greatly reduced, and in addition, the grid voltage would have to be much smaller to avoid distortion (Vol. I).

(b) The oscillatory current flows through the H.T. battery, giving rise to back-coupling and instability.

The above discussion of various types of power stage should also help to emphasise the importance of having the impedance of the output circuit in the correct ratio to the A.C. resistance of the valve system. This is not always easy to arrange, but the process is considerably simplified if transformer coupling to the output stage is used. The effective output impedance then depends on the transformation ratio, thus introducing another adjustable factor (*cf.* N.56).

32. **Use of Controllable Reaction in Amplifiers.**—In all amplifiers, especially R/F amplifiers, reaction, or back coupling, is present to a certain extent, either intentionally or unintentionally.

In dealing with regenerative amplification, in Section "D," we had an illustration of the intentional use of **inductive reaction** to diminish the damping of the input circuit of a receiver, and hence to increase the strength of signal heard in the telephones. In that case it was observed that excessive reaction led to continuous oscillations, the reaction receiver being, in fact, the basis of a transmitter (K.3).

Various other circuit arrangements can be used to give effectively the same result.

Instead of an inductance a capacity may be used to give reaction in the way shown in Fig. 18 (*a*). The R/F anode current I_a has an alternative path back to filament through the variable condenser and grid-filament tuned circuit. The proportion of current which takes this path depends

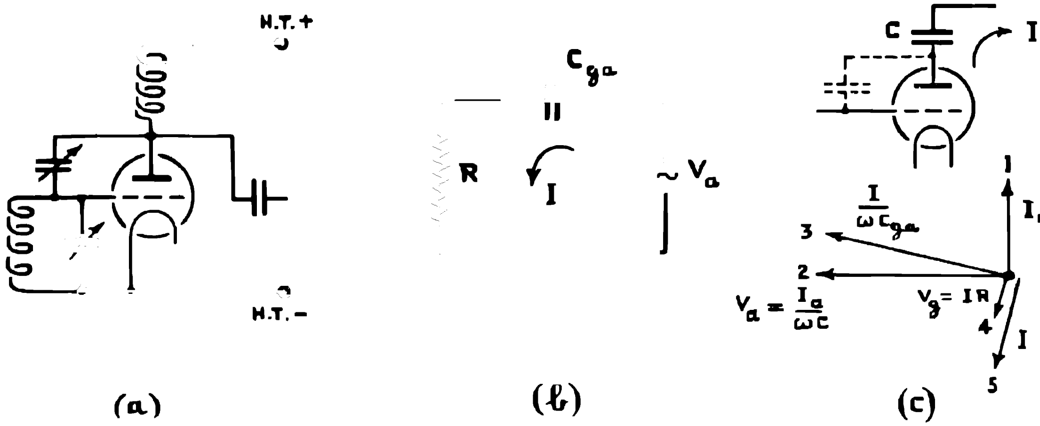


FIG. 18

on the reactance of the condenser, and so is controllable. It may be considered that this artificial condenser increases the effective value of the grid-anode inter-electrode capacity, since it is connected in parallel; the nett reactance through this path is thereby reduced, and the process is known as **capacitive reaction.**

In general, the energy fed into the grid circuit in this way may either assist or oppose the original input current according to the phase of the fed back current. A full discussion of this point is complicated but has been treated by Miller and others, from which the name **Miller effect** has originated; only the result can be given here.

For practical purposes it may be taken that **unless the output circuit has nett capacitive reactance [Fig. 18 (c)], the energy fed back in this way is such as to assist the original oscillation in the input circuit.** With a capactive output, however, the feed back damps the original oscillations; in effect it constitutes a load on the input circuit and increases its positive resistance. This matter can be given a simple qualitative explanation by means of the vectors of Fig. 18 (*c*), the numbers on the vectors representing the order in which they may conveniently be studied (*cf.* Section "D," and K.4, 6). With reference to the anode current I_a the anode oscillatory potential is V_a, and is drawn 90° *leading* since it has the nature of a back E.M F. The latter is to be regarded as the applied E.M.F. in the equivalent circuit of Fig. 18 (*b*), consisting of C_{ga} and a resistance R representing the resonant impedance of the tuned grid circuit. With this simplification vector (3) represents the voltage across C_{ga}, and vector (4) represents the voltage drop across the resistance R; the current vector I must be in phase with vector (4). From this it is evident that the effect of reaction is the production of an oscillatory input between grid and filament tending to give an anode current more than 90° out of phase with the anode current which caused the action. Moreover, the diagram shows that V_a and V_g are *less* than 90° out of phase; in K 6 it is shown that a circuit tends to be self-oscillatory (possessing negative resistance) when V_a and V_g are *more* than 90° out of phase. From two points of view, therefore, the circuit appears to be not of the self-oscillatory type, but rather the reverse; the output impedance constitutes a damping load (N.29)

Controllable reaction is of great importance and use, but many of the troubles of wireless are due to unintentional reaction which may sometimes be uncontrollable. Controllable reaction is further discussed at a later stage.

Theoretically, the damping of the tuned circuit may be reduced to zero and hence the amplitude of the oscillation increased to infinity. In practice, it is necessary to leave a margin of stability, so that the amplifier may not be too near the border line at which it generates self-oscillations. Light damping gives highly selective circuits.

33. **Self-Oscillation in Amplifiers.**—Especially at high frequencies and with multi-valve amplifiers, it is more usually the case that sufficient reaction, inductive, capacitive or resistive, is inherent in the amplifier to give a tendency to self-oscillation without the use of intentional reaction.

Reaction coupling of these various types enables a portion of the oscillatory energy in circuits near the output side of the amplifier, where this oscillatory energy assumes large dimensions, to be handed back to circuits near the input side of the amplifier, tending to nullify their positive resistance.

Inductive coupling may occur if inductances used in the amplifier are not carefully screened from one another.

Capacitive coupling may occur by means of stray capacities between leads.

A much more important reason for the occurrence of capacitive coupling between circuits is, however, the presence of capacity between the different electrodes of the valves used in the amplifier, a matter which has led to the use of tetrodes and pentodes.

The most important inter-electrode capacity, from this point of view, is that between grid and anode, usually denoted by C_{ga}.

If we take the simple case of a tuned circuit between grid and filament and a tuned anode output circuit, as illustrated in Fig. 19, it is easy to see that they are coupled together by virtue of C_{ga}, and therefore energy can be fed from the tuned anode circuit back to the input circuit. In

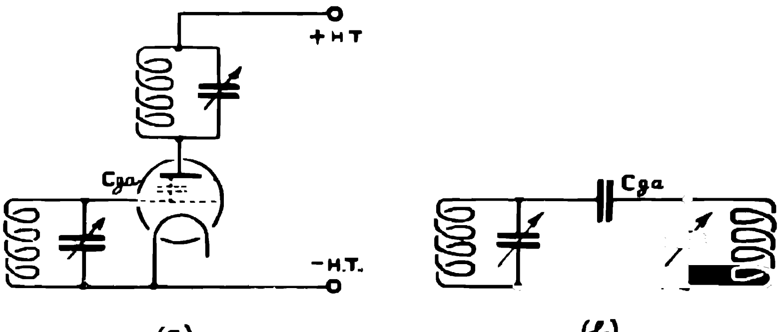


FIG. 19.

fact, this type of coupling, in this case undesirable, is exactly the same as the deliberate capacitive coupling illustrated in Fig. 18, and if the output circuit is suitable, as is generally the case in amplifier stages, energy is fed back to the input circuit.

The greater the value of C_{ga}, the higher is the coefficient of coupling between the circuits, and the greater the amount of energy fed back.

Part of this inter-electrode capacity is, of course, between the lead-in wires to the electrodes, and this can be arranged to be small by keeping the wires as far away from each other as possible.

Resistive coupling may occur by circuits being coupled through the resistances of the batteries and common leads to these. Oscillations in the anode circuit of one valve cause high frequency voltage variations across these resistances, which are transferred to the anode circuits of other valves fed from the same source.

These various types of accidental reaction coupling may lead to self-oscillation where self-oscillation is not wanted, *e.g.*, when spark, I.C.W. or R/T are being received, since these are already modulated at the transmitter.

If the amplifier is designed to receive C.W. as well, it is necessary to have some arrangement by which a local oscillation is produced, either by separate heterodyne or by sufficient reaction being available and **controllable**, so that the amplifier may act as an autodyne when necessary.

It is the problem of unwanted oscillation which sets a definite limit to the amplification obtainable with any model, the difficulty of preventing it increasing with the number of valves used, the V.A.F. which results, and the frequency which is being amplified.

The fact that, with the usual output circuits, the tendency to self-oscillation increases with the number of valves, and the amplification produced, can easily be shown as an extension of the case of capacitive coupling by anode-grid capacity illustrated in the last paragraph.

If we take the case of a multivalve amplifier instead of a single valve amplifier, the output circuit of each valve being tuned, then Fig. 20 represents a simplified arrangement of the circuits and the capacities which couple them together.

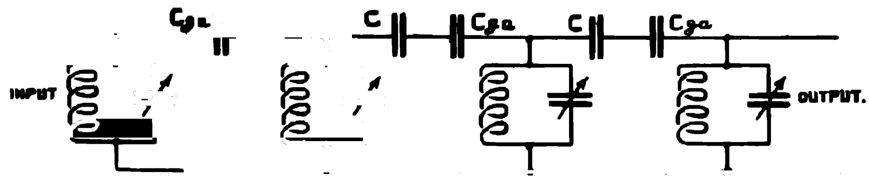


FIG. 20.

The coupling capacities are the inter-electrode capacities C_{ga} and the insulating condensers in the grid circuits of the valves.

Now with several valves in cascade, as in this case, the last output circuit is coupled to the input circuit of the first valve by a capacity which varies inversely as the number of valves.

Neglecting the insulating condensers C, which are large compared with C_{ga}, and, therefore, do not materially affect the equivalent capacity between each pair of circuits, we have in the above figure a number of capacities C_{ga} in series.

If there are n stages of amplification, the equivalent value of all of these is $\frac{C_{ga}}{n}$, which, as stated above, varies inversely as n.

But the amplification increases in geometrical progression with the number of stages, and so the amplitude of oscillatory energy in the last tuned circuit increases correspondingly. The amount of energy feed-back is proportional to the energy in the output circuit multiplied by the value of the coupling capacity, and as the output energy increases much faster than the coupling capacity decreases, their product must also increase when the number of steps is increased.

Hence, any amplifier will generate self-oscillations if the number of stages is increased sufficiently, because the total energy fed back is always increasing, and must finally be greater than the damping losses in the input circuit.

34. **Prevention of Self-oscillation in Amplifiers.**—Various methods are adopted for dealing with the tendency of amplifiers to oscillate, and these will now be described in detail. They may be classified as follows :—

(1) **Simple precautionary measures,** designed to prevent inductive and capacitive coupling between different portions of the circuit.

Such are :—

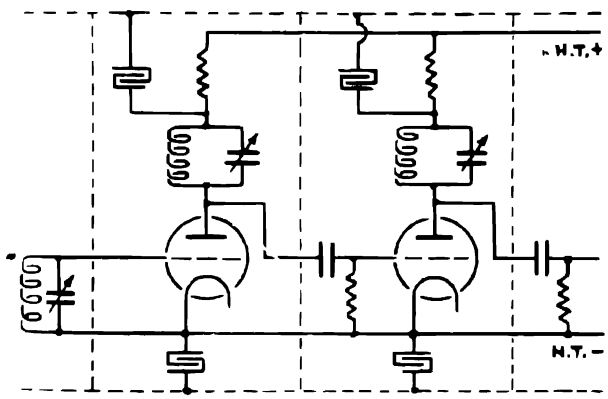
Use of short leads, well spaced apart, crossing at right angles.

Separation of the stages of the amplifier from each other by earthed metal screens. This leads to a loss of energy, through eddy currents induced in the screens.

(2) **By-pass condensers** across batteries and potentiometers to avoid resistive coupling.

Large condensers shunted across batteries or potentiometers present to R/F currents only a small reactance. Without these, the resistance of the batteries or potentiometers would cause considerable oscillatory voltage variations across them. With common H.T. supply, this P.D. would be a source of energy to every stage, whilst in the case of a potentiometer every valve whose grid was connected to it would be affected. In the case of batteries, the effect would be increased as they were used up, owing to their increasing internal resistance.

With efficient amplifier stages, it is found that this precaution is not enough to prevent feed-back, since the coupling due to the common lead from the H.T. battery to the anodes of the various valves, and to the common filament leads, is sufficient to produce self-oscillation. It is necessary in modern amplifiers to conduct the radio frequency anode current by the shortest possible path to filament once it has passed through the output impedance. This is achieved by the use of **de-coupling condensers and resistances,** as illustrated in Fig. 21. The oscillatory anode current, after passing through the tuned anode output circuit, has two paths in parallel open to it, one through the de-coupling condenser and the other through the resistance in the anode lead. The reactance of the condenser is much less than the resistance, and so the anode current is almost completely by-passed to the screen, to which each filament is also connected by a large condenser.



DECOUPLING

FIG. 21.

(3) **Reduction of filament current or H.T. voltage.**

These reduce the V.A.F. of any stage by reducing the amplification factor of the valve, and so the energy fed back is also diminished.

(4) **Reaction applied deliberately "the wrong way."**

A reaction coil or condenser may be connected so as to prevent the generation of oscillations instead of assisting it. This will be the case if an ordinary reaction coil has its connections reversed.

In some amplifiers the reaction coil or condenser can be connected to various anodes in the amplifier. **It can be shown that if, with the connection made to the odd anodes, oscillations tend to be maintained, the joining up of the connection to an even-numbered anode will tend to stop oscillations.**

This arises from the fact that in any one valve of the amplifier, the oscillatory grid and anode voltages are more than 90° out of phase, *e.g.*, with resistance-capacity coupling they are 180° out of phase, since the anode current and grid voltage are in phase while the anode current and anode voltage are 180° out of phase. The anode of one valve is connected to the grid of the next through the coupling condenser. It follows, therefore, that the grid voltages of two succeeding valves in an amplifier are more than 90° out of phase, and similarly for the anode voltages.

Though this always holds for two successive valves in an amplifier, caution should be exercised in applying it, for instance, to the first and last stages of an amplifier. Due to various reasons, the anti-phase relationship is not strictly preserved. It can easily be seen, for example, that in a choke-capacity-coupled amplifier, where even in successive valves corresponding voltages are not in exact anti-phase, the rule given above for applying "positive" or "negative" reaction is not justified.

Negative reaction obviously reduces amplification.

(5) **Increase of resistance in the input circuit.**

One method of doing this is simply to increase the ohmic resistance.

This obviously acts as a deterrent to oscillation, making the amount of energy fed back less capable of overcoming the damping losses.

Another method is to bias the grid positively so that grid current flows. This is equivalent to saying that the A.C. resistance between grid and filament is decreased from its theoretically infinite value (when the grid is negative) to some finite value, and the condenser of the input circuit has, therefore, a finite resistance connected across it. By the same theory as that used in Vol. I such a resistance is equivalent to a series resistance in the oscillatory circuit itself, which is greater the less the value of the shunt resistance. Hence the total effective resistance of the input circuit is increased, and the energy fed back is less capable of overcoming the damping losses.

It should be noted that if the positive grid bias is so large that it causes saturation grid current to flow, the grid-filament A.C. resistance again becomes infinite, and produces no damping of the input circuit.

In practice, the grid is biased positively by taking the filament connection from the input circuit to the sliding contact of a potentiometer across the filament battery (*cf.* paragraph 25 and Fig. 12). It is usually only necessary to maintain the first grid at a positive potential. Obviously, as with previous methods, the increased damping of the input circuit results in a loss in amplification.

(6) **Flattening the input and output circuits.**

It is found that flattening either of these circuits increases the stability of an amplifier.

The V.A.F. is, of course, decreased if the output circuit is flattened $\left(\frac{L}{CR}\text{ becomes less}\right)$, and, while flattening the grid circuit does not affect the V.A.F., it does mean a diminution in input voltage across the condenser of the tuned input circuit, for a given voltage induced into it from the aerial, and therefore results in less total amplification from aerial to output.

Methods (1) and (2) deal with specific causes of reaction ; the others make it more difficult for reaction present in the amplifier to produce undesired effects.

The above methods give a reduction of amplification as a necessary accompaniment to greater stability.

The following measures, which need not reduce the amplification, are therefore more efficient.

(7) **Neutralising Circuits.**—Neutralising consists in the insertion of a condenser or condensers to produce an equal and opposite reaction effect to that inherent in the amplifier by reason of the inter-electrode capacity between grid and anode. Before the invention of the screen-grid valve the process was widely applied both in receiver construction and in transmitters. Since the year 1929, however, we have witnessed a change in the former, and the passing of the neutralised triode with the growth in popularity of screen grid and R/F pentode valves ; the use of neutralisation is now almost restricted to transmitters (K.39).

There are many ways of joining up such neutralising condensers. Two typical circuits are illustrated below, known in the Service as the **tapped output** and **tapped input** methods respectively

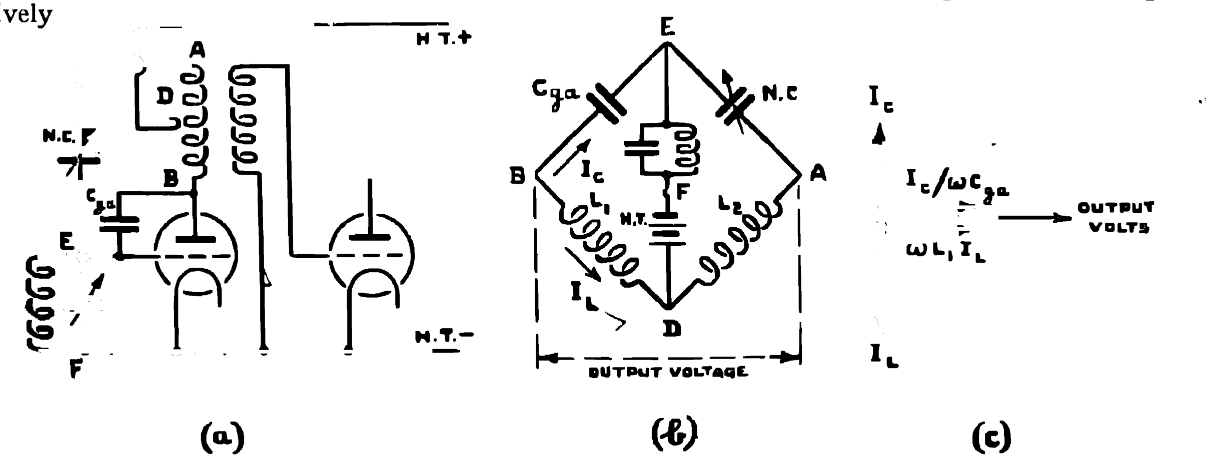


FIG. 22.

In Fig 22 (*a*), the H.T. **is fed to the valve at the mid-point of the inductance in the anode circuit,** which may be a choke coupling, or the inductance arm of a tuned-anode coupling, or the primary of a transformer ; it is called **the tapped output method.**

From the other end of the inductance a neutralising condenser (N.C.) is joined up as shown.

The energy conveyed through C_{ga} and N.C. from the output circuit is fed into the grid circuit from opposite ends of the output coil, the mid-point of which is at approximately the same oscillatory potential as the filament. The P.Ds. across the two halves of the output coil are therefore 180° out of phase, as far as the input circuit is concerned, and so the supplies of oscillatory energy through the two coupling condensers C_{ga} and N C. are in opposite phases, and tend to neutralise each other. Neutralisation of the inter-electrode capacity feed-back is therefore obtained by adjusting the variable neutralising condenser until the feed-back through it is exactly equal to that through C_{ga}.

The circuit can be redrawn as an A.C. bridge system. The theory of balancing such a **Wheatstone bridge** for direct currents was explained in Vol. I. The same condition holds for *impedances* in the A.C. case as for *resistances* in the D.C. case; the system is, however, complicated by the fact that the **voltages in the balancing arms must be equal not only in magnitude but also in phase.**

The neutralising circuit in this form is shown in Fig. 22 (*b*). As regards feed-back, the source of energy is the output circuit, which therefore corresponds to the battery in the D.C. case. It will be seen that, if the bridge is balanced, no current due to the output voltage can flow through the input circuit. The operation of neutralising thus consists of adjusting the variable condenser

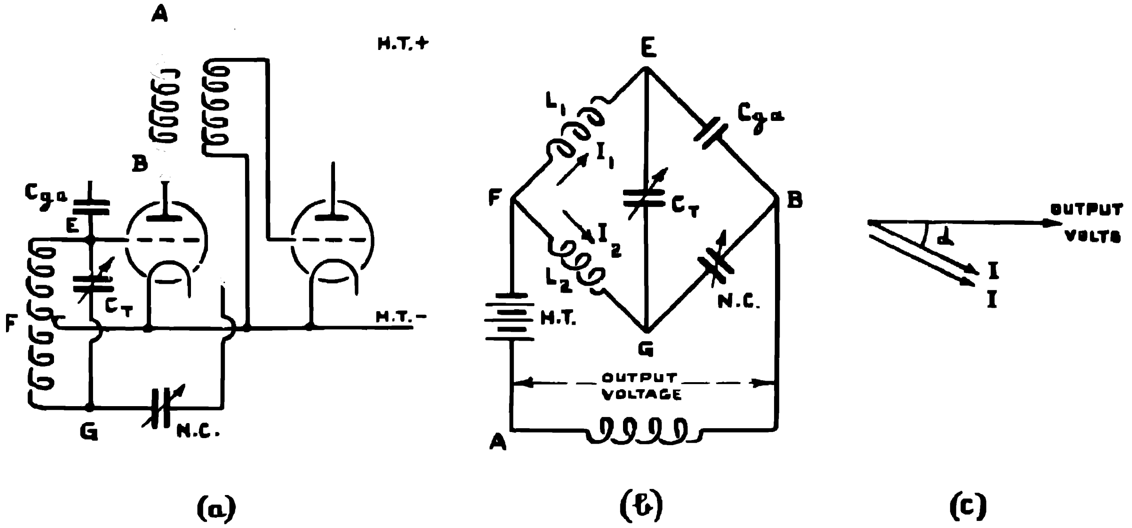


FIG. 23.

C. until this condition is satisfied. In the circuit as drawn, this would be the case if the capacity of N.C. were equal to C_{ga}, since D is supposed to be at the electrical mid-point of AB. In practice, owing to the mutual flux-linkage between the two halves of AB, and various stray capacities, it is impossible to arrange the tapping point D so as to fulfil accurately the above condition.

The chief practical result is that the correct setting of the neutralising condenser varies with the frequency received (K.40, 41).

The balancing may be mathematically expressed as follows :—

$$\omega L_1 \; I_L = \frac{I_1}{\omega C_{ga}}$$

$$\omega L_2 \; I_L = \frac{I_2}{\omega C_{NC}}$$

Hence
$$\frac{L_1}{L_2} = \frac{C_N}{C_{ga}}$$

From this it follows that the currents in the arms BEA and BDA need not be equal in order to achieve the balance. A vectorial representation is given in Fig. 22 (*c*). With reference to the output volts applied between B and A, the current in BEA will lead, and that in BDA will lag, each by 90°. Under these conditions, the oscillatory voltage across C_{ga} may be in phase with and equal

in magnitude to that developed across inductance L_1. If L_1 equals L_2 then the output volts will be given by a vector twice the length of $\omega L_1 I_1$.

Another common neutralising circuit and its equivalent bridge are shown in Fig. 23. In this case the centre of the input coil is tapped to filament, and so only **half the input voltage is applied between grid and filament.** The behaviour of this circuit will be obvious from the figure and the explanation of the previous neutralising circuit. The capacity of N.C. is adjusted until the currents flowing in the bridge arms, due to the output voltage, produce the same oscillatory P.Ds. across N.C. and C_{ga}. No energy can then be fed into the input circuit; this is called **the tapped input method.**

With reference to the output volts, the currents in the arms FEB and FGB will lead or lag depending upon whether the nett reactance is capacitive or inductive respectively. Assuming the nett reactance to be inductive in each case, the current I may lag on the output volts by some angle α; so long as two current both lag by the same angle, it is possible to adjust the oscillatory voltages across L_1 and L_2 to the requisite equality for a balance. If the currents are also equal (I), a balance will result when L_1 equals L_2; this is the case represented in Fig. 23 (*c*).

In general, we have

$$\omega L_1 I_1 = \omega L_2 I_2 \quad \therefore \quad \frac{L_1}{L_2} = \frac{I_2}{I_1}$$

$$\frac{I_1}{\omega C_{ga}} = \frac{I_2}{\omega C_{NC}} \quad \therefore \quad \frac{C_{NC}}{C_{ga}} = \frac{I_2}{I_1}$$

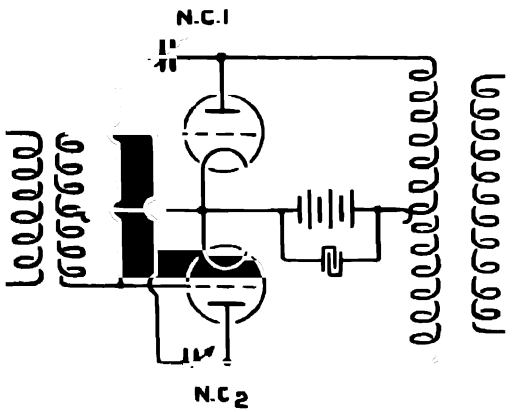
$$\therefore \quad \frac{L_1}{L_2} = \frac{C_{NC}}{C_{ga}}.$$

It may be remarked that the variable condenser N.C., in the above circuits, instead of being adjusted strictly for neutralisation may be used to control the amount of energy fed back through C_{ga}, *i.e.*, the amount of energy fed back through N.C. may be less than the regenerative feed-back through C_{ga}, provided the former is large enough to prevent self-oscillation. In this case N.C. may be regarded as a **negative reaction capacitive coupling.**

If more energy is fed back through N.C. than through C_{ga}, the damping of the input circuit will actually be increased beyond that due to its own losses. This will tend to occur as the capacity of N.C. increases above C_{ga}.

It will be seen, however, that N.C. (and half the input inductance in series with it) is in parallel across anode and filament with the output circuit. Considerable increase of N.C. may thus eventually result in this total output circuit having a nett capacitive reactance. With a capacitive output circuit, the feed-back through C_{ga} damps the input circuit, while that through N.C. is regenerative, and there is therefore a possibility of self-oscillation.

A neutralising condenser is never likely to be large enough for this to happen, but damping of the input circuit by feed-back through C_{ga} often occurs in detector valves. The output circuit is designed, for example, to have an inductive reactance at audio frequencies, but its self-capacity may be such that at radio frequencies the nett reactance of the coil is capacitive. Thus the feed-back of radio frequency energy to the input through C_{ga} is in such a direction as to damp the input oscillation and reduce the efficiency of the receiver (paragraph 32).



NEUTRALISED PUSH-PULL CIRCUIT.

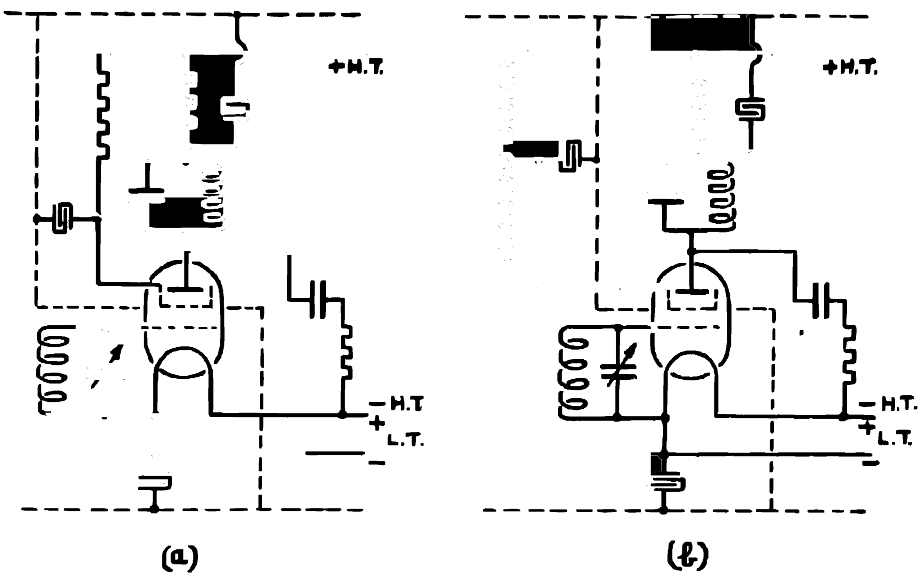
FIG. 24

(8) **Push-Pull Circuits.**—It has already been pointed out (paragraph 27) that with properly-matched valves in a push-pull arrangement, and mid-point grid bias, the oscillatory current flowing through the H.T. battery and common leads may be reduced to an exceedingly small amplitude and the risk of resistive coupling due to this cause thereby diminished. Owing to its

symmetrical nature, the push-pull arrangement also lends itself readily to the use of neutralising condensers for balancing inter-electrode capacitive coupling. Such a neutralised circuit is shown in Fig. 24. The neutralising condensers are connected between the anode of one valve and the grid of the other. It will be seen that this system combines the advantages of both the previous neutralising circuits given for one valve, both input and output being centrally tapped.

(9) **Four-Electrode (Screen Grid) Valve, or Tetrode.**—The characteristics of this valve were fully discussed in B.41, and it was pointed out that the inter-electrode capacity C_{ga} between grid and anode could be reduced to an exceedingly small value. **Thus the use of such valves provides a powerful method of cutting down the feed-back from output to input** through this capacity, and hence of preventing instability.

To obtain full advantage from this very loose internal capacitive coupling, it is essential that the external coupling between output and input should also be reduced to a minimum, *i.e.*, the external screening between grid and anode circuits must also be as complete as possible (*cf.* B.16, 41). The use of metallised valves has considerably simplified this process. In the case of Fig. 25 the external screening is shown by dashed lines.



SCREEN GRID STAGE IN AMPLIFIER.
FIG. 25.

Erratum.—Condensers in tuned anode circuits should be variable.

Fig. 25 (*a*) and (*b*) illustrates respectively two methods of obtaining the screen voltage from the H.T. supply.

(*a*) A resistance is inserted between H.T. positive and the screen terminal. The screen current flowing through this resistance thus produces a P.D., and the screen potential is less than that of H.T. positive by the amount of this P.D. Thus, to quote some usual figures, the screen current might be 0·5*m*A with a 50,000 ohms resistance. The P.D. across the resistance would then be 25 volts, and, with 120 volts H.T., the screen potential would be 95 volts.

(*b*) **A tapping is taken to the screen from a high resistance potentiometer.**

It should be noted that the screen circuit constitutes an impedance Z_2 forming a common coupling between the grid and anode circuits, as shown in Fig. 26. With an oscillatory input between grid and filament there will, of course, be an oscillatory screen current, just as there is an oscillatory anode current. With a finite external impedance in the screen grid circuit this oscillatory current would set up oscillatory potentials on the screen which would constitute a coupling between the grid and anode circuits partially invalidating the action of the screening grid. If the screen circuit presents a negligible impedance, the oscillatory variation of screen voltage will also be negligible, this is achieved by connecting the screen to filament and earth through a large condenser, presenting a negligible impedance at the frequency in use. Thus, **as regards oscillatory potential the screen and filament may be considered common,** although there is, of course, a large constant P.D. between them.

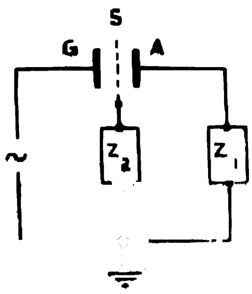


FIG. 26.

35. **Tetrode Amplification.**—It was seen in Section "B" that the amplification factor m and A.C. resistance r_a of the screen grid valve were very high as compared with those of the triode. This is not primarily because higher values of these constants could not be obtained with triodes, but because no advantage could be taken of the higher amplification possible, owing to the instability it would certainly produce through inter-electrode capacity coupling. The very small amount of the latter in the tetrode allows a stable amplification stage to be used, while giving a **much higher V.A.F.**

The theoretical calculation of the V.A.F. is the same as for the triode. Apart from its effect on the constants, the extra electrode does not affect the algebra of the valve. Thus, with an external resistance R in the anode lead, the V.A.F., as before, is V.A.F. $= \dfrac{mR}{R + r_a}$.

The disadvantage of resistance-capacity coupling, from the point of view of the extra H.T. voltage necessary, has already been discussed for triodes. In tetrodes, the much larger value of r_a requires a corresponding increase in the external resistance, to take advantage of its amplifying properties, and so this disadvantage is still more pronounced In general, therefore, **a tuned anode output circuit** will be employed, acting effectively at its resonant frequency, as a very high resistance. In practical cases, the A.C. resistance of the valve is much higher than that of the tuned circuit connected to the anode, and R is almost negligible in comparison with r_a. Under these conditions an approximate formula for V.A.F. may be used; by substituting for m we have V.A.F. $= g_m$R. This shows that for high amplification we need a "high slope" valve followed by a tuned circuit of high impedance (*cf.* B.44).

The superiority of screen grid valve amplification may best be seen by a numerical illustration, taking from the average values quoted in Section "B" the following constants for triode and tetrode :—

—	g_m	r_a	m.
	mA/volt	Ohms	
Triode	0·75	20,000	15
Tetrode	0·5	200,000	100

With an effective external resistance of 100,000 Ω in each case, the V.A.Fs. are

(*a*) Triode :—

$$\frac{15 \times 100{,}000}{120{,}000} = 12\cdot5.$$

(*b*) Tetrode :—

$$\frac{100 \times 100{,}000}{300{,}000} = 33\cdot3.$$

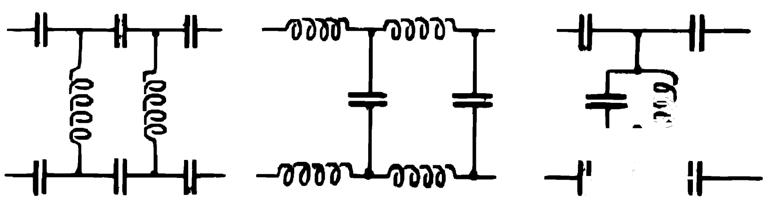
The advantage of the tetrode is obvious, and further, with the figures quoted, the tendency to instability would be much less with the tetrode than with the triode. It can also be shown that tetrode amplification is **more selective.**

Since tetrodes prevent feed-back from the output circuit to the input circuit they frequently act as **isolating valves** in the first stage of an R/F amplifier, their primary function being to minimise re-radiation of energy from the aerial.

Tetrodes are **unsuitable in A/F stages** where large input voltages are applied between grid and filament. The reason for this is explained below.

36. **Pentode Amplification.**—In the discussion on dynamic characteristics in B.39 it was shown that the slope of the operating characteristic of a triode with a load in the anode circuit is much less than the static slope. This is because a rise in anode current, brought about by a change in grid voltage, will produce an increased drop in volts across the impedance of the external anode circuit, with a corresponding fall in the actual potential on the anode. This fall in anode volts is responsible for a reduction of anode current, so that the latter does not rise to the extent which the static characteristic would indicate as a result of the given change in grid volts. Now an examination of the anode characteristics of the screen grid valve, Fig. 31, Section "B," shows that for a wide range of anode voltage above the screen potential the curve is nearly horizontal. In other words, fairly large changes of anode voltage have little effect on the anode current, and therefore the slope of a dynamic mutual characteristic is not much less than that of the static curve. Thus this type of valve practically retains its high amplification factor and mutual conductance when in operation. This property is very desirable for audio frequency amplification, but the screen grid valve is limited for this purpose because the dip or kink in the anode characteristics, due to secondary emission effects, limits the available grid swing to half a volt or so. The introduction of the earthed grid in the pentode suppresses the effects of secondary emission, and extends the range of grid swing, without distortion of anode current, to 30 volts or more, making the construction an excellent one for the power output stage of an amplifier (*cf.* B.42).

37. **Filter Circuits.**—The term "Filter Circuit" is used loosely to cover any type of circuit used in conjunction with an amplifier to remove undesired frequencies, and thus give selectivity. The subject is a very specialised one and is not fully treated here.



FILTER TO PASS HIGH FREQUENCY (a)

FILTER TO PASS LOW FREQUENCY (b)

BAND PASS FILTER (c)

FIG. 27

If it is desired to eliminate audio frequency interference the " **high pass** " circuit shown in Fig. 27 (*a*) may be used. Here, radio frequency currents will pass readily through the condensers, while audio frequency currents will find the condensers a path of high reactance, and will prefer the short-circuit path through the inductances.

Fig. 27 (*b*) indicates the opposite type of circuit, namely, one designed to pass audio frequency or direct current, and impede the passage of radio frequency current ; it is a " **low pass** " circuit.

If designed to pass direct current only, as when joined up in the D.C. source of supply to a valve receiver, the inductances will have iron cores, and the condensers will have values of the order of microfarads.

The circuit shown in Fig. 27 (*c*) includes a rejector circuit tuned to a particular frequency. Current at this and adjacent frequencies will therefore be unable to pass through it, and must pass along the line as required.

Currents at other frequencies will find an easy short-circuit path through the rejector.

It is called a " **band-pass filter,** " as it separates a band of frequencies in the neighbourhood of its resonant frequency from other frequencies which were originally present. Its frequency response curve may be compared with that of two tuned circuits mutually coupled.

The filter must be terminated by the correct impedance, which is determined by the ratio of the capacities and inductances used.

38. **Note Filter Circuit.**– A "Note Filter" or " **Note Selector** " circuit is used to add a further degree of selectivity, by allowing **audio-frequency** currents of one particular frequency only to pass through the telephones.

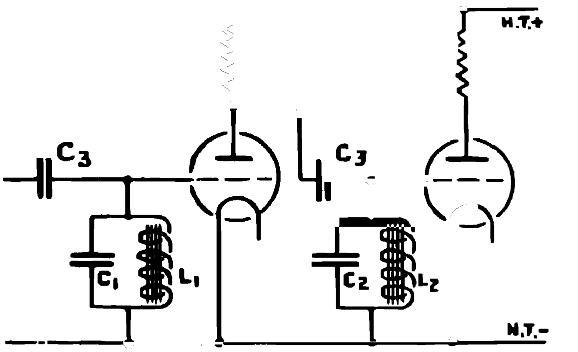


FIG. 28.

This application is of great importance. It makes possible the operation of several different channels of communication on the same radio frequency, using I.C.W. with different modulating frequencies (*cf.* paragraph 59).

Fig. 28 illustrates a typical circuit. It represents a combination of a note magnifier circuit, with rejector circuits L_1C_1, and L_2C_2.

The heterodyne in the radio frequency stage of the receiving gear is adjusted to give a note (in combination with the required C.W. signal) of the same frequency as that to which the rejector circuits are tuned.

This signal will therefore suffer no loss in the rejectors, and will be magnified and passed through the telephones in the usual manner.

C.W. signals of other frequencies, however, giving other notes in combination with the heterodyne oscillation, will find a short-circuit path through the rejector circuits, L_1C_1 and L_2C_2, and should not reach the telephones at all.

39. **A Modern Straight Circuit (T.R.F. Type) Service Receiver.**—Fig. 29 represents the electrical details of a modern receiver in which the components, aerial isolating stage, tuner, R/F amplifier, detector, heterodyne oscillator, and A/F amplifier, are all easily discerned (paragraph 25). The instrument was designed to cover a portion of the L/F and M/F ranges, and to be operated from either A.C. or D.C. cathode heating supplies.

Two alternative methods of connecting the aerial to the set are provided. In one case, the aerial is coupled by a differential condenser (1) to the grid of a screen grid isolating valve (2). The output from this valve is fed across a transformer coupling to the tuner, the H.T. supply to the valve being decoupled in the usual manner. This arrangement effectively isolates the aerial from the tuned circuits of the receiver, and permits a number of receivers to be used on the same aerial without mutual interference. The differential condenser maintains constant the total capacity in the aerial circuit, and effectively acts as an input volume control.

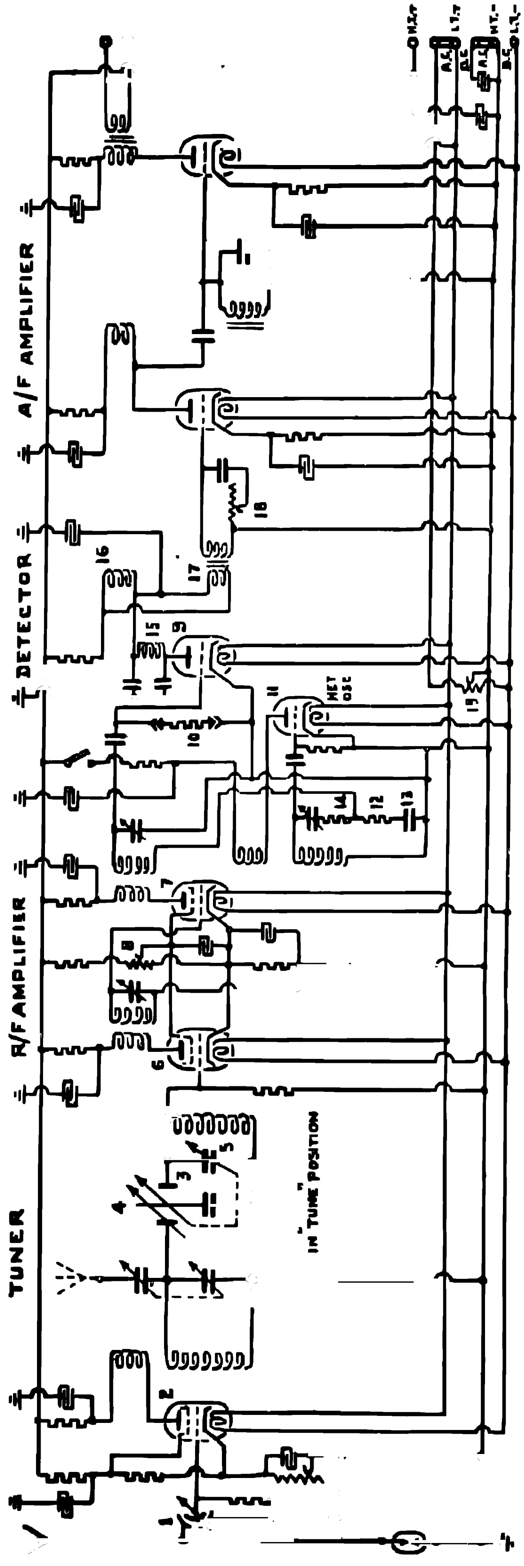


FIG. 29.

Erratum.—***Output of detector*** ; ***de-coupling condenser*** *shown connected between coils* 16 *and* 17 *should be* ***connected to top end*** *of* 16.

In the second case, where isolation of the aerial is not desired, the aerial is plugged directly into the tuner.

The tuner consists of a primary and secondary circuit, coupled by means of a variable "top-capacity coupling" (D.45 and Vol. I), having a compensating device giving constant selectivity. The diagram shows the set in the "tune" position, but arrangements may be made to have a less selective "stand-by" position, in which the primary circuit is eliminated and tuning is effected by the secondary circuit only, the latter being directly coupled to the aerial.

The coupling condenser (3) is of some interest, the fixed and moving plates consisting of half cylinders, the inner or moving plate being adjustable, by an external dial, in order to vary the coupling capacity. Between these two plates a cylindrical *earthed* screen (4) is made to slide in an axial direction by a lever which is actuated by a cam attached to the secondary tuning condenser spindle (5). The coupling capacity depends on the extent to which the earthed screen is interposed between the condenser plates. If it were fully interposed there would be no capacity and the two circuits would be completely uncoupled. It is observed in D.45 that capacitive couplings tend to produce over-coupling, with consequent loss of selectivity, as the frequency increases; in this case, over coupling is avoided by arranging automatically to decrease the coupling capacity as the frequency increases. The shape of the cam is such that the effective coupling capacity varies with the tuning of the secondary circuit, to give practically constant selectivity over the whole range of the receiver. At the same time, the selectivity can be varied by adjusting the moving plate of the condenser by an external dial.

It will be noted that the cylindrical screen (4) is *earthed* through a condenser, the value of which is 1 jar, *i.e.*, 1,000 centimeters. This condenser is very large in comparison with the two halves of the top-capacity coupling (3), which have maximum values of 10 centimetres and 20 centimetres respectively, reading from left to right in the diagram. The coupling between the circuits provided by the 1-jar condenser connected to the screen will therefore be very small in comparison with that attributable to top capacity. The two circuits are so loosely coupled that it is possible to assume that one circuit does not appreciably affect the tuning of the other; this enables one to arrive at a simple numerical estimate of the coupling due to the 1-jar condenser. The tuned primary circuit may be considered to be connected to a capacity potentiometer, consisting of the 10 centimetre and the 1,000 centimetre condensers joined in series; the P.Ds. available across the latter two condensers will be in the ratio of 1,000 : 10. The 1-jar condenser may now be regarded as a source of E.M.F. attached to a capacity potentiometer consisting of a 20 centimetre condenser and the 1,000 centimetre condenser, which is part of the secondary circuit; the P.Ds. available across the latter two condensers will be in the ratio of 1,000 : 20. Very *approximately*, the coupling between primary and secondary circuits is therefore given by the expression

$$K = \frac{10}{1000} \times \frac{20}{1000} = \frac{2}{10^4} = 0 \cdot 02 \text{ per cent.}$$

This coupling is small, even in comparison with the smallest value of coupling factor provided by the top-capacity coupling, the latter assumed to be acting by itself. In that simple case the coupling factor is given by the formula quoted in D.44; the minimum value of coupling capacity is of the order of 3 centimetres, and hence the coupling factor is given approximately by

$$K = \frac{3}{\sqrt{1000 \times 1000}} = 0 \cdot 3 \text{ per cent.}$$

For the above reasons, one may ignore any coupling due to the 1-jar condenser connected to the screen, regarding the latter as being effectively earthed.

The output from the tuner is applied between grid and cathode of the first R/F screen grid valve (6). The R/F stages (6) and (7) are transformer coupled, with secondaries tuned, the secondary circuits thus forming the tuned grid circuits of the succeeding valve.

The H.T. supply for the screens of the two R/F valves is controlled by a potentiometer (8) which acts as a volume control; the usual decoupling arrangements are provided.

The detector valve (9) operates on the cumulative grid principle, the grid leak (10) being inserted between grid and cathode when using A.C. indirectly heated valves.

The heterodyne oscillator (11) is of the familiar reversed feed type, employing a tuned grid circuit and reaction coil in the anode lead. The heterodyne frequency is mixed with the incoming signal frequency by a form of grid injection (F.51), the 4-ohm resistance (12) and the 0·5 μF condenser (13) being common both to the oscillator grid circuit and the tuned grid circuit of the detector valve. The resistance and condenser combination presents a high impedance at low frequencies, and therefore gives a strong heterodyne coupling; as the frequency increases the impedance of the combination decreases, and provides the requisite weak heterodyne coupling on the higher frequencies. The 100-ohm resistance (14) prevents serious interference with the tuning of the heterodyne and detector grid circuits due to their being coupled.

The output from the detector valve passes through a two-stage low-pass filter (15), having a cut-off at 7,000 cycles. Audio-frequencies below 7,000 cycles are passed on to coil (16), across which is connected in parallel the transformer primary (17) which serves as the coupling to the A/F amplifier.

The note magnifier consists of two A/F stages which may be used as note magnifiers or note selectors or a combination of both, the diagram showing the arrangement for one stage of note selection followed by note magnification.

The note selector stage employs the principle of parallel-feeding (*cf.* N.49), the output impedance consisting of a tuned grid circuit, which is transformer coupled (17) and connected in parallel with the anode choke (16). The tuned grid circuit is sharply tuned to 1,350 cycles per second, using a 7-henry inductance tuned by a fixed condenser of 1-jar and a 0·5-jar variable one. The selectivity of the tuned circuits may be made so great that a slight "ringing" occurs when a signal is received. This is not sufficient to prevent the reading of morse at hand operated speed, but becomes a disadvantage when high speed automatic reception is necessary. A selectivity control (18) is therefore used which introduces resistance into the tuned grid circuit, reducing the decrement from about ·05 to 0·3; damping is introduced with consequent reduction of selectivity. The transformer coupling also serves to render inappreciable the damping of the tuned circuit due to the anode-filament impedance of the preceding detector valve; for this purpose the primary of the transformer has a smaller number of turns than the secondary, a very high step-up ratio being used. This is equivalent to a very weak coupling, the resistance reflected from the secondary circuit to the primary being very small and given by R/T^2, where R is the secondary resistance and T is the transformer ratio. By reciprocal action the effect of the primary on the secondary is equally small, and may be considered equivalent to a very large resistance in parallel with the tuned circuit. The voltage amplification of the stage due to transformer step-up effect is about 1 : 1.

The second A/F stage is shown in the note magnifier position. Parallel-feeding is employed for the tuned grid circuit constituting the output impedance, the latter employing the same components as in the preceding stage. In this case, however, the circuit is highly damped, due to the replacement of transformer coupling by choke-capacity-coupling; the anode-filament impedance of the previous valve is, in effect, connected in parallel with the tuned grid circuit which has a very much larger impedance at resonance. As a result of this, the resonance of the tuned circuit is so flattened that the stage is unselective, and gives a fairly even response from about 200 to 2,000 cycles per second.

When the circuit is arranged for two stages of note magnification the first A/F stage is similar to the one described above.

Other features of interest include the provision for using either D.C. valves or A.C. indirectly heated valves, in which respect the diagram is considered self-explanatory; there is also the hum potentiometer (19), a "hum dinger," the operation of which is described in N.63. Automatic grid bias (N.41) is obtained in certain cases by means of resistances inserted in the cathode leads, suitably by-passed by condensers.

40. **H/F Amplification.**—Owing to the difficulty of designing suitable output circuits, the methods of R/F amplification so far described become inefficient at high frequencies, and different methods must be employed.

In this section, references to H/F amplification are intended to be used in a general sense to cover the whole band of intermediate, high and very high frequences (*cf.* Prefatory note on Nomenclature of Waves).

41. **Difficulties of H/F Amplification.**—The general methods of amplification already discussed in this section (with the exception of *reaction*), depend for their success on the possibility of designing an output impedance which is comparable with, or, if possible, greater than the A.C. resistance of the valve. This is comparatively simple for low and medium frequencies and the lower intermediate frequencies, but is rendered impossible at higher frequencies by the presence of unavoidable capacities, which by-pass the high-frequency currents.

The self-capacity of the coil or resistance used as output impedance, the anode-filament capacity of the valve and the grid-filament capacity of the next valve, are all in parallel across the output of any amplifier stage, and so the total output impedance cannot be made greater than the combined reactance of these three capacities. An average value for this reactance at 20,000 kc/s., say, would be about 5,000 ohms, which is, of course, much less than the A.C. resistance of any valve used for radio-frequency amplification. This difficulty in obtaining efficient amplification applies equally both to triodes and tetrodes, as may be illustrated by calculating the amplification obtained with an output impedance of 5,000 ohms for the two valves considered in paragraph 35.

The V.A.Fs. are :—

(*a*) Triode—

$$\text{V.A.F.} = \frac{15 \times 5{,}000}{25{,}000} = 3.$$

(*b*) Tetrode—

$$\text{V.A.F.} = \frac{100 \times 5{,}000}{205{,}000} = 2 \cdot 44.$$

The amplification obtained is thus very small in both cases. Further, in the case of a triode amplifier its external impedance is likely to have a nett capacitive reactance on the higher frequencies which it is designed to receive. When this occurs, the feed-back from the output to the input circuit through C_{ga} is such as to damp the input and reduce the total amplification still more. On the other hand, if the output reactance is not capacitive, the tendency to self-oscillation will be strong, owing to the large regenerative feed-back through C_{ga}, the reactance of which decreases as the frequency increases.

42. **Methods of H/F Amplification.**—Of the methods of radio frequency amplification already discussed, there remains only one, viz., amplification by the use of **reaction,** which is suitable for H/F reception. In addition to the T.R.F. type there are available two general types of receiver which have not yet been mentioned. These are :—

(i) **Super-heterodyne receivers.**
(ii) **Super-regenerative receivers.**

The ordinary methods of note magnification after detection are obviously still available.

It should be noted that the super-heterodyne receiver is also widely used over the L/F and M/F wave bands in both Service and commercial practice.

43. **General Features of H/F Receivers.**—Before proceeding to the discussion of methods of amplification, some general points about high frequency receivers may be noted.

(1) **Untuned aerials** are generally used. High frequency receivers, as will be seen below, normally contain self-oscillatory circuits, and, unless special precautions are taken, enough radiation may take place with a tuned aerial to cause interference with other sets in the neighbourhood. This radiated energy also increases the damping of the receiver, and is likely to be variable, as the unavoidable small changes in the natural capacity of the aerial vary its natural frequency. This renders the operation of the receiver spasmodic, and may, on occasion, damp the self-oscillatory stage so much that oscillations cease altogether.

(2) When building a receiver to work over a considerable range of frequencies, it is not always a simple matter to design an aerial whose natural frequency lies outside the required range. For his reason the **coupling between the aerial circuit and the first valve circuit is made as oose as possible,** the necessary amplification being obtained in later stages. This may be achieved by resistance coupling, or the use of a coupling condenser of small capacity. Another common method, which may be used in conjunction with the above, is to insert as first stage a valve circuit which give little or no amplification, but serves as a further device to **isolate the self-oscillatory stage from the aerial.** A screen grid valve is more efficient for this purpose than a triode, since the coupling between output and input circuits by inter-electrode capacity is considerably less.

(3) To avoid the introduction of self-capacitive by-pass paths, which are inevitable when inductances are tapped for tuning purposes, plug-in coils are used to cover the different frequency ranges. The winding of these can be arranged to **minimise self-capacity effects** over the range for which they are designed.

(4) As far as possible, **the use of long leads is avoided.** The impedances of leads may be quite appreciable at high frequencies, and, instead of acting as good conductive paths, they may display considerable opposition to the passage of high frequency currents. To reduce their impedance to a small value, large condensers may be inserted between the leads and earth. These are commonly known as " high frequency earths."

44. **Reaction Receivers for H/F Work.**—The principles underlying the use of *controllable* reaction to increase the strength of incoming signals have already been considered, and various amplifiers embodying it have been described in this section and Section " D." Another common type of reaction is illustrated in Fig. 30. This may be described as **capacity-controlled magnetic reaction.** The amount of energy fed back to the input circuit via the coupling coil is controlled by adjustment of the variable condenser. This renders control of the amount of reaction much smoother than in the case of simple **inductive reaction** obtained by varying the adjustment of a swinging coil in the anode lead; moreover, the tuning of the LC circuit is not subject to the alterations necessarily imposed by the " swinging coil " method. The valve in Fig. 30 will normally be the detector valve, since earlier amplifying stages are ineffective, and it will be seen that there are two parallel paths from anode to filament in the output circuit, one through the reaction circuit and the other through a choke. The current through the detector valve contains both radio and audio frequency components; the choke presents a relatively low impedance to audio frequency oscillations, but a high impedance to radio frequencies; the reaction circuit acts in exactly the opposite manner, and so the radio and audio frequency output components are separated. The **radio frequency component** returns to filament via the reaction circuit. It is thus all usefully employed, and, further, is kept out of the audio frequency stages. The **audio frequency component** passes on through the choke to the telephones or first note magnifier stage, as the case may be.

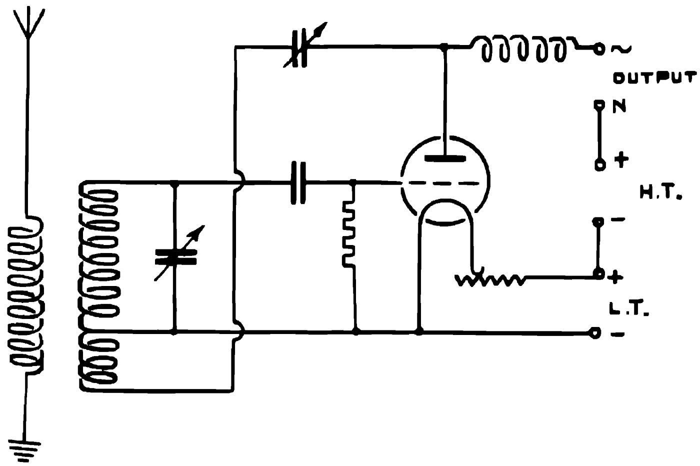


FIG. 30.

Reaction receivers give the best results, from the point of view of signal strength and selectivity, when reaction is increased close up to the point where self-oscillation takes place. There is the

disadvantage that any local disturbance such as irregularities in filament emission, filament vibration, and battery noises, are also magnified. There is thus a background of extraneous noise in the telephones, known as "mush," in addition to the note produced by the incoming signal.

In modern receivers, the reaction control is usually sufficiently sensitive to permit close approach to the oscillating point, without actually causing self-oscillations. In some of the older receivers the control was coarse, and receivers were often operated in an unpleasant oscillating condition which seriously distorted the note on any incoming signal.

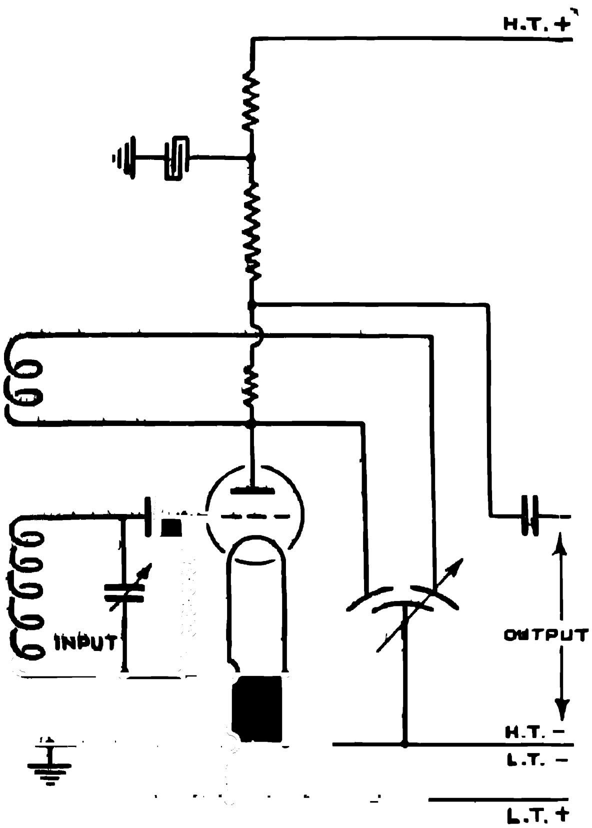


FIG. 31.

Inductive reaction is described in Section "D," simple capacitive reaction in paragraph 32, and another form of capacity-controlled magnetic reaction in paragraph 46, Fig. 33. Fig. 31 represents an improved form of capacity-controlled magnetic reaction employing a differential condenser; the latter varies the capacitive impedance in the reaction circuit, but maintains constant the total capacity in parallel with the reaction coil, so that alteration in the amount of reaction does not greatly alter the tuning. Here again, the valve is the "detector," and the A/F output is shown to be resistance-capacity coupled to the next stage.

45. **Threshold Howl in Reaction Receivers.**—When the detector valve of a reaction receiver employing cumulative grid detection is followed by a transformer-coupled stage of note magnification, it may happen that reaction is increased too much, self-oscillations commence, and a low frequency oscillation is also produced, which is evident in the telephones as a continuous note. This phenomenon is known as "threshold howl." The note rises in pitch as the reaction is increased. As mentioned above, greatest sensitivity is obtained by operating reaction receivers just below the oscillating point, so that the occurrence of threshold howl is particularly possible and undesirable. It becomes really serious in the case of the autodyne reception of C.W., since it occurs just at the most sensitive point for oscillation and detection.

The production of this howl is due to the audio-frequency change in anode current on which detection depends. In cumulative grid detection this change is a decrease (Section "D"), and the better the conditions for detection, the greater will it be. The anode current flows, of course, through the external inductance in the anode lead and so a back E.M.F. is set up in this inductance. When the anode current is decreasing, this back E.M.F. is in such a direction as to assist the H.T. battery, and therefore increases the voltage applied to the anode. As the amplitude of the high-frequency self-oscillations is proportional to the anode voltage, it follows that these oscillations build up to a larger amplitude than would be the case if the external inductance were absent. When the anode current reaches its lowest point and is momentarily steady, the back E.M.F. disappears and so the amplitude of the self-oscillation decreases.

As the anode current rises again to its initial value, the back E.M.F. will be in the reverse direction, *i.e.*, opposing the H.T. battery. Thus the anode voltage is still further decreased and the amplitude of the self-oscillations falls correspondingly. When the anode current steadies and the

anode voltage is simply that due to the H.T. battery, the amplitude of self-oscillation rises again and the cycle of events is repeated. The result is that the amplitude of the high-frequency oscillations is **modulated** at an audible frequency, and a howl is heard in the telephones. Fig. 32 is a possible picture of the oscillation.

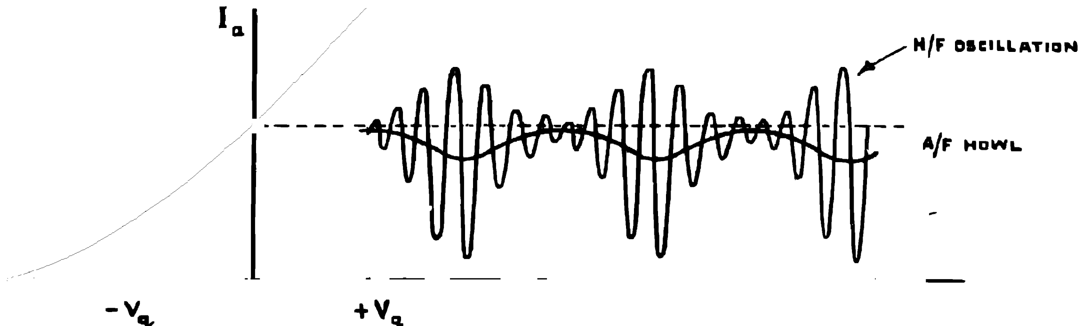


FIG. 32.

A similar effect would tend to be produced in an anode bend detector with a condenser in the high frequency output circuit, as in this case both the audio frequency change in anode current and the back E.M F. across the condenser are in the opposite direction.

In many cases this audio frequency swing is rapidly damped out, and the effect does not attain the dimensions of a "howl." In a circuit of this kind, however, transient damped oscillations, similar to the type described as "mush" above, tend to be accentuated and to increase the background of noise in the telephones. The tendency towards audio frequency self-oscillations by back coupling with the later stages of note magnification is also encouraged.

The inclination to howl is generally only pronounced when the output choke is of large inductance, and there is a large audio frequency drop in anode current. For instance, the choke used to separate the R/F and A/F components in the circuit shown in Fig. 30 may be large enough to carry out its function without necessarily setting up threshold howl.

The rise in pitch of the note with increasing reaction is due to the decrease in the time constant $\frac{L}{R}$ of the circuit, where L is the output inductance and R is the total series resistance through which the audio frequency current flows. This resistance, since it includes the valve A.C. resistance, increases as the mean grid potential runs more negative, which will be the case with increasing reaction because of the increased amplitude of the oscillatory grid voltage. Hence $\frac{L}{R}$ decreases as the reaction increases, and the pitch of the note rises.

The simplest method of obviating threshold howl in a reaction receiver is to modify the nature of the output impedance. In the case considered above, a purely inductive output produces audio frequency variations of anode voltage in the correct phase to the anode current variations to maintain self-oscillation, and this phasing is fairly critical. The phasing conditions for the maintenance of these **A/F oscillations** are the same as for any self-oscillator ; the oscillatory anode volts should be as nearly as possible in anti-phase with the oscillatory grid volts (*cf.* K.6). There is no possibility of threshold howl with an output circuit employing resistance-capacity coupling to the next stage (paragraph 32).

It is possible to preserve, to some extent, the greater amplification of transformer coupling, while preventing threshold howl, by shunting the transformer primary by a resistance. This alters the phase of the audio frequency P.D. developed across the output, and therefore the anode voltage variation, sufficiently to prevent self-oscillation. The value of the resistance can be kept high enough, however, to give sufficient audio frequency current through the transformer primary to obtain greater amplification than would be the case if a pure resistance-capacity coupling were used.

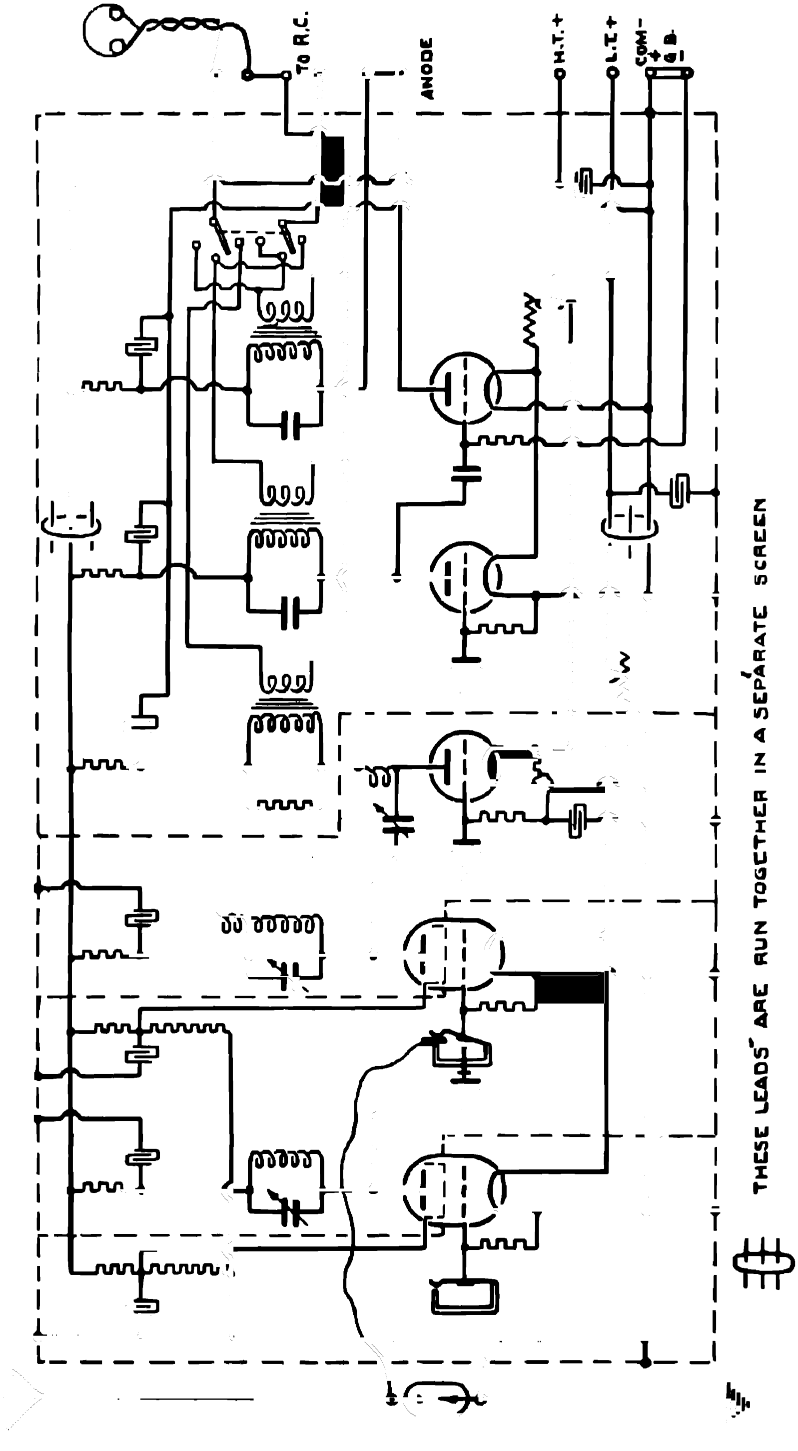

To R.C.
ANODE
H.T.+
L.T.+
COM-
+ G.B. -
THESE LEADS ARE RUN TOGETHER IN A SEPARATE SCREEN

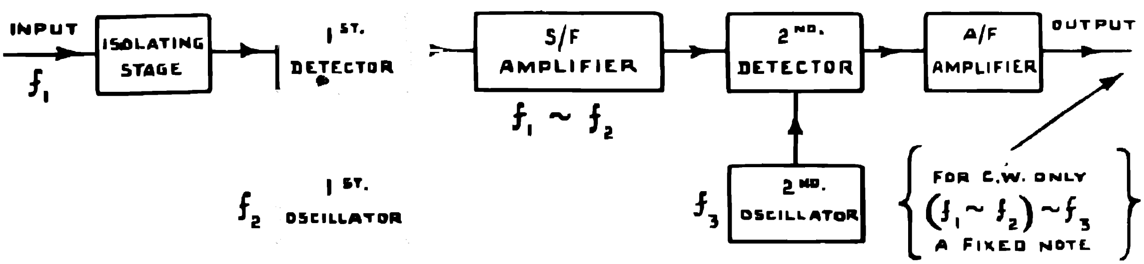
FIG. 33.

The grid bias in a cumulative grid detector may also be altered from its best value for detection, *i.e.*, made more negative. Obviously this also reduces efficiency, but the smaller audio frequency decrease in anode current lessens the tendency to oscillation.

46. **Typical Service Reaction Receiver.**—A modern Service receiver employing reaction and designed to cover the I/F and H/F bands is shown in Fig. 33. The aerial is untuned, and is loosely coupled to the first stage by resistance coupling. Two stages of tuned anode radio frequency amplification using screen grid valves are provided, but only give efficient amplification on the lower frequencies. One or both stages may be employed, and a plug and jack fitting is provided on the grids for this purpose. At higher frequencies the first stage is not used, and the second stage is mainly useful as an isolating valve. The third valve is a cumulative grid detector, the appropriate grid bias being obtained by a tapping from a potentiometer across the filament terminals. Capacity-controlled magnetic reaction is employed in this stage. One, two or three stages of note magnification may be used as shown. The output choke in the first audio frequency stage gave rise to a tendency to threshold howl, which was counteracted by adjusting the grid bias of the detector, as mentioned above, and by shunting the choke with a large resistance. The careful screening, and the de-coupling condensers and resistances should also be noted.

47. **Principles of Super-heterodyne Receivers.**—Invented in 1917, the general principle of these receivers is to heterodyne the incoming H/F signal with a locally-generated H/F oscillation, the frequency difference between the two oscillations being arranged to lie well above the audible limit, *i.e.*, to be *supersonic* and of low radio frequency The resultant oscillation is rectified, and the low frequency component can then be efficiently amplified at a convenient frequency, by the ordinary low frequency methods, before final detection and note magnification. The advantages of radio frequency amplification are thus preserved, while avoiding the difficulties incidental to its application at the frequency of the incoming signal. This use of amplification at a frequency intermediate between that of the incoming signal and an audible frequency gives this circuit its name of super-heterodyne, or **supersonic heterodyne receiver.** The method is convenient and efficient, and is widely used, even at those lower frequencies where the ordinary methods of amplification can also be employed. The amplification at a fixed frequency in the intermediate stages simplifies tuning adjustments considerably.

Fig. 34 is a schematic diagram of a superheterodyne tuner amplifier. The input R/F signal of frequency f_1 passes through an isolating stage to the first detector. The first local oscillator supplies an input of frequency f_2 ; a fixed supersonic beat frequency $f_1 \sim f_2$ is passed on to the S/F amplifier.



SCHEMATIC DIAGRAM OF A SUPERHETERODYNE TUNER AMPLIFIER

FIG. 34.

Following this stage, with modulated signals such as R/T, is a second detector and A/F amplifier. In the case of C.W. signals an audible result can only be obtained with the help of a second oscillator, adjusted so that the beat is within the audible range.

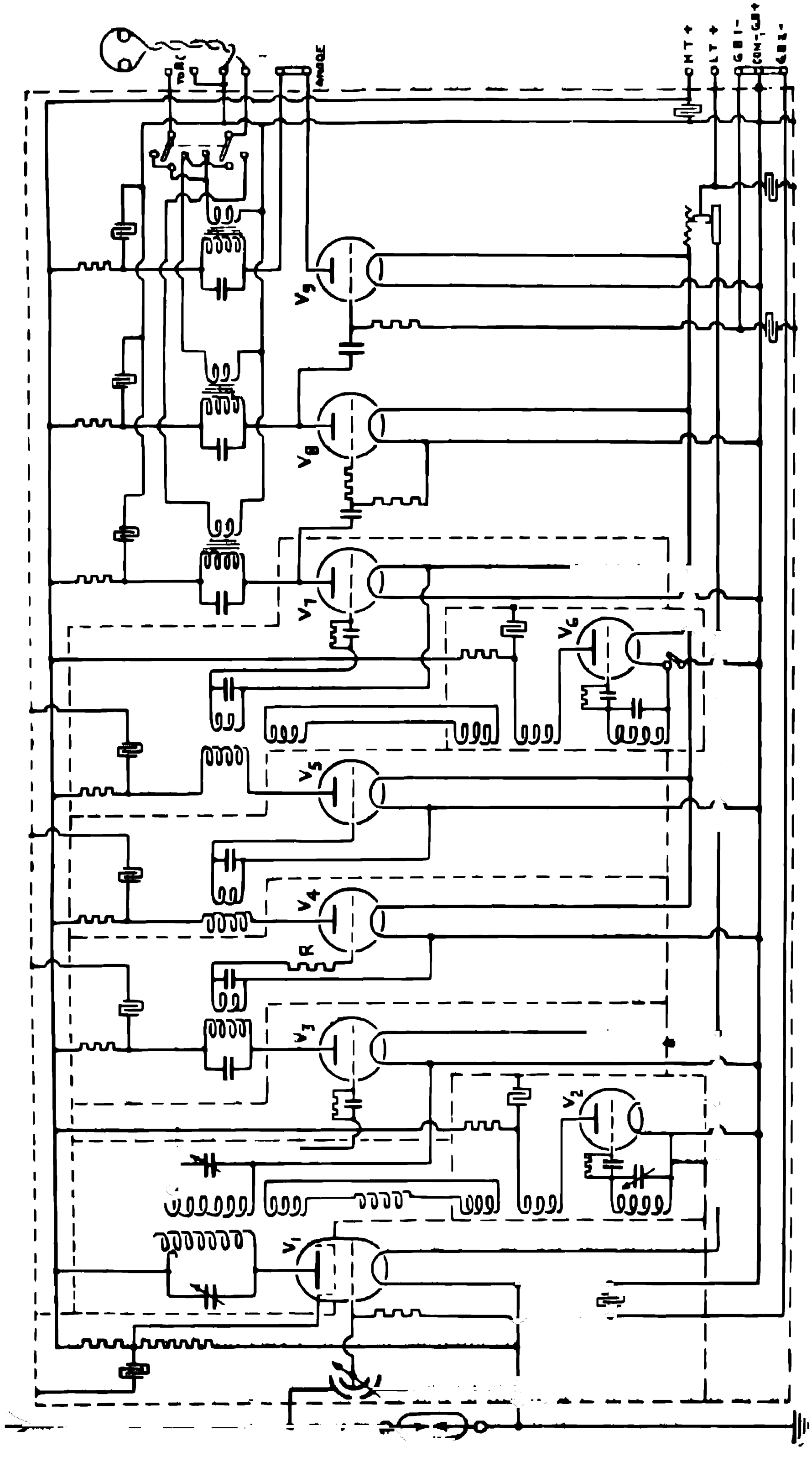


Fig. 35.

48. **A Modern L/F Superheterodyne W/T Receiver.**—The principle will now be further discussed with reference to a modern Service superheterodyne receiver shown in Fig. 35. The aerial, as before, is untuned, and the first R/F stage is a screen grid valve with tuned transformer output, mainly included to act as an **isolating stage.**

The aerial input is applied by means of a **differential condenser** which, while maintaining constant the capacity to earth in the aerial circuit, acts effectively as an **input volume control.** The differential condenser consists essentially of two condensers in series, and rotation of the movable plate alters the ratio of the capacities of these condensers, the nett capacity remaining constant. With powerful input signals, the input oscillatory volts between grid and filament may be reduced by decreasing the reactance of the lower condenser of the combination. To do this the moving plate must be brought to the lower position to cover the lower fixed plate ; the lower capacity will then be a maximum and the reactance a minimum. An input control of this nature is necessary, since, otherwise, powerful signals might suffer partial rectification at this stage.

V_2 is the first local oscillator, and is coupled to the grid circuit of V_3 by an untuned mutual link circuit, all these precautions being designed to minimise coupling to the aerial as much as possible. The combined oscillations due to the incoming signal and local oscillator are applied between the grid and filament of V_3, which is a cumulative grid detector. The beat frequency is 30 kc/s. in this model. The high frequency component of the output is mainly by-passed to the screen, and the 30 kc/s. oscillation, which is of constant amplitude if C.W. is being received, but is modulated at audio frequency in the case of an I C.W. signal, is passed on to the grid-filament circuit of V_4 by the transformer coupling shown The resistance R in the grid lead of V_4 is known as an **R/F stopper.** Any residual R/F potential difference developed at the terminals of the transformer secondary is applied across R and the grid-filament capacity C_{gf} in series. At high radio frequencies the resistance of R is much greater than the reactance of C_{gf}, so that only a small proportion of the total P.D. is applied between grid and filament. At supersonic frequencies the reverse is the case, and so R acts as a device for cutting down R/F input voltages without sensibly affecting S/F input voltages. V_4 and V_5 provide two stages of amplification at 30 kc/s., using tuned transformer coupling. The amplified oscillation is detected in V_7, the method of cumulative grid rectification again being employed. V_8 and V_9 are stages of note magnification. The resistance in the grid lead of V_8 performs a similar function to that in the grid lead of V_4, filtering S/F and A/F potential differences. When I.C.W. is being received, the incoming signal is modulated at audio frequency, and so an audio frequency note is obtained by detection in V_7. In the case of C.W., however, provision must be made for heterodyning the 30 kc/s. oscillation so as to produce an audio frequency note. This is accomplished by means of another separate local oscillator V_6, as shown ; the tuning of the latter is fixed.

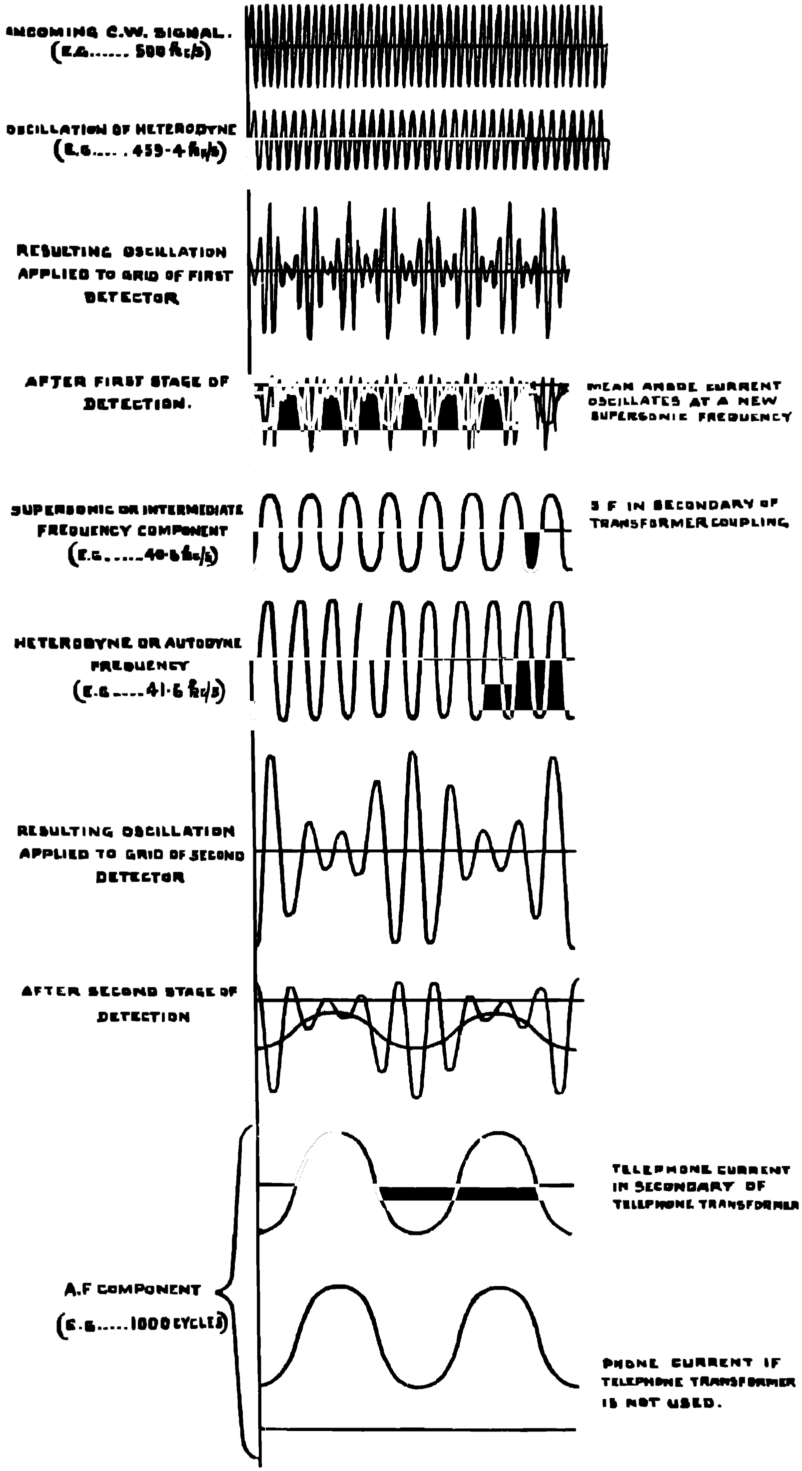
It will be seen that the only tuning adjustments required are in the output circuit of V_1, the first local oscillator V_2, and the grid circuit V_3, so that in spite of its apparent complexity the circuit is simple to operate.

The curves of Fig. 36 and Fig. 37 give a pictorial representation of the process taking place at each stage in any superheterodyne receiver.

49. **Selectivity of the Superheterodyne Receiver. Choice of S/F.**—The superheterodyne circuit is very selective, and this is one of the chief reasons for extending its use down to the low frequency range, where the less elaborate circuits of the earlier portion of this section can give efficient amplification.

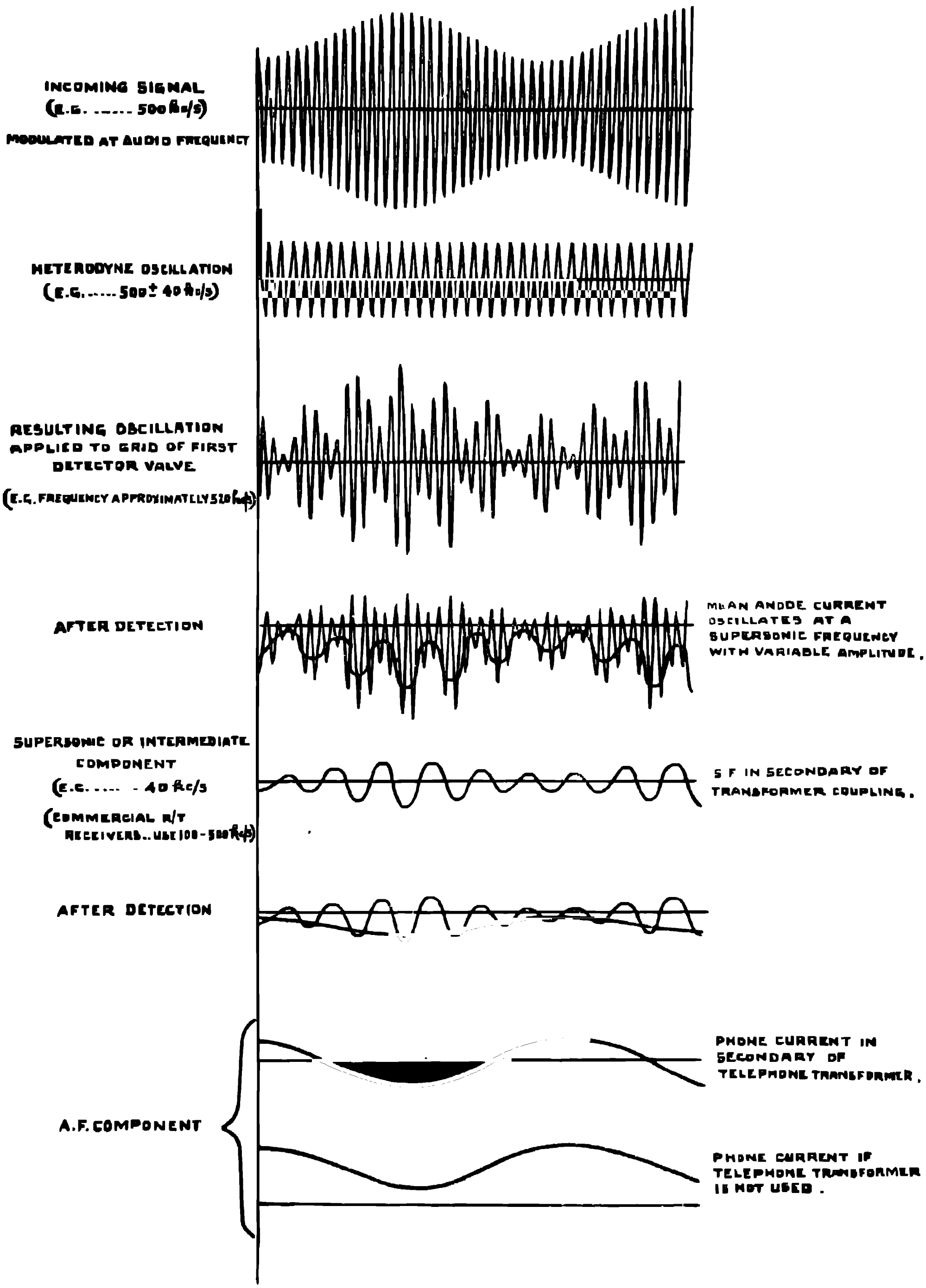
Suppose we wish to receive a signal on 1,000 kc/s., using a heterodyne frequency of 960 kc/s., in order to give an S/F of 40 kc/s. Any **adjacent channel interference** on (say) 990 kc/s. would produce a beat note of 30 kc/s., and would easily be stopped by the I/F circuits tuned to 40 kc/s., since it is 25 per cent. off resonance. In this way the interfering signal is very much weakened compared with the wanted signal by the time it reaches the second detector valve ; the lower the S/F, the greater the adjacent channel selectivity.

It is important, however, to note that although the adjacent channel selectivity of a receiver with a low I/F is very high, there are two incoming frequencies at which a given setting of the first local oscillator will give the same beat frequency. Considering the above numerical example, an



RECEPTION OF C.W. BY SUPERHETERODYNE RECEIVER.

FIG. 36.



RECEPTION OF A MODULATED WAVE BY A SUPERHETERODYNE RECEIVER

FIG. 37.

unwanted signal on 920 kc/s. and a *wanted* one on 1,000 kc/s. will both give a note beat of 40 kc/s. Interference of this kind is known as **second channel interference**, the elimination of which necessitates pre-selector circuits capable of discriminating between 1,000 kc/s. and 920 kc/s.; this is not too easy a matter, since there is a percentage difference of only 8 per cent. Where the first oscillator cannot be separately controlled, this type of interference can only be avoided by a careful choice of the S/F beat frequency, having regard to the interfering frequencies most likely to be encountered. In cases where the first oscillator is controllable, considering the above numerical example, the practical method would be to change the heterodyne frequency of the first oscillator to 1,040 kc/s. in an attempt to eliminate the interference.

It is interesting to consider the effect of changing the intermediate frequency to some higher value such as 400 kc/s. In this case we will suppose that we wish to receive a signal on 1,000 kc/s. using a heterodyne frequency of 600 kc/s. to give a 400 kc/s. beat note. The possibility of second channel interference on 200 kc/s. need hardly be considered, since it could easily be eliminated by the tuned circuits. The situation is not so favourable, however, in the case of adjacent channel interference on say, 990 kc/s.; the latter would produce a beat note of 390 kc/s., and it will be difficult to eliminate that in S/F stages tuned to 400 kc/s.

The choice of the intermediate frequency varies with the use to which the superheterodyne receiver is to be put, or with the type of selectivity which it is desired to achieve. The simpler commercial domestic R/T receivers are required to possess a high degree of freedom from "second channel interference"; since the frequency of the first oscillator is usually not under separate control, this is most easily arranged by using a relatively high intermediate frequency (from 100 to 500 kc/s.), the values 125 kc/s. and 450 kc/s. being commonly found. A high degree of protection against adjacent channel interference is not usually required. In the case of receivers used mainly for W/T purposes, separate control of the first oscillator is usually provided, and second channel interference is usually avoidable by placing the heterodyne frequency above or below the wanted signal, as described above; to provide a high degree of protection against adjacent channel interference relatively low intermediate frequencies of the order of 40 kc/s. are used, 20 kc/s. being about the limiting value.

In cases where these receivers are required to work only in the H/F band, satisfactory adjacent channel selectivity can be obtained with proportionately higher intermediate frequencies; for working on 25,000 kc/s an S/F of 120 kc/s. may be used (*cf.* paragraph 50). In commercial receivers, values up to 500 kc/s. may be found.

50. **A Modern H/F Superheterodyne W/T Receiver.**—Fig. 38 represents, diagrammatically, the electrical details of a receiver of more advanced type than that shown in Fig. 35; with suitable care it is capable of giving high amplification and selectivity from 3,000 kc/s. up to frequencies of the order of 23,000 kc/s

The receiver is divided into six screened compartments containing, respectively, the aerial tuner, R/F amplifier, first detector, intermediate frequency amplifier together with the second detector and A/F amplifier, the first heterodyne oscillator, and the second heterodyne oscillator.

The aerial tuner is a closed oscillatory circuit in which the inductance is made in two halves with a space between them, and with right and left-hand windings symmetrically disposed about the centre point which is connected to the case and earth terminal. In the central space is a coupling coil, consisting of two turns wound in opposite directions and capable of moving between the two halves of the aerial inductance; this coil forms a portion of tuned grid circuit of the R/F amplifier stage. The object of the unit is to enable the input impedance of the receiver to be varied in order to match approximately the impedance of any aerial system which might be used; the right and left-hand windings give an additive effect from the anti-phase input obtainable from an aerial system balanced with respect to earth.

The aerial inductance is provided with three pairs of symmetrical tapping points to which the two aerial terminals A and B can be connected by flexible leads and plugs, assuming that the aerial is a dipole aerial, or double array with feeders balanced to earth. When the aerial circuit is tuned, the impedances between the pairs of tapping points are approximately 100 ohms, 500 ohms, and 2,500 ohms respectively, half these values representing the impedance between one tapping point

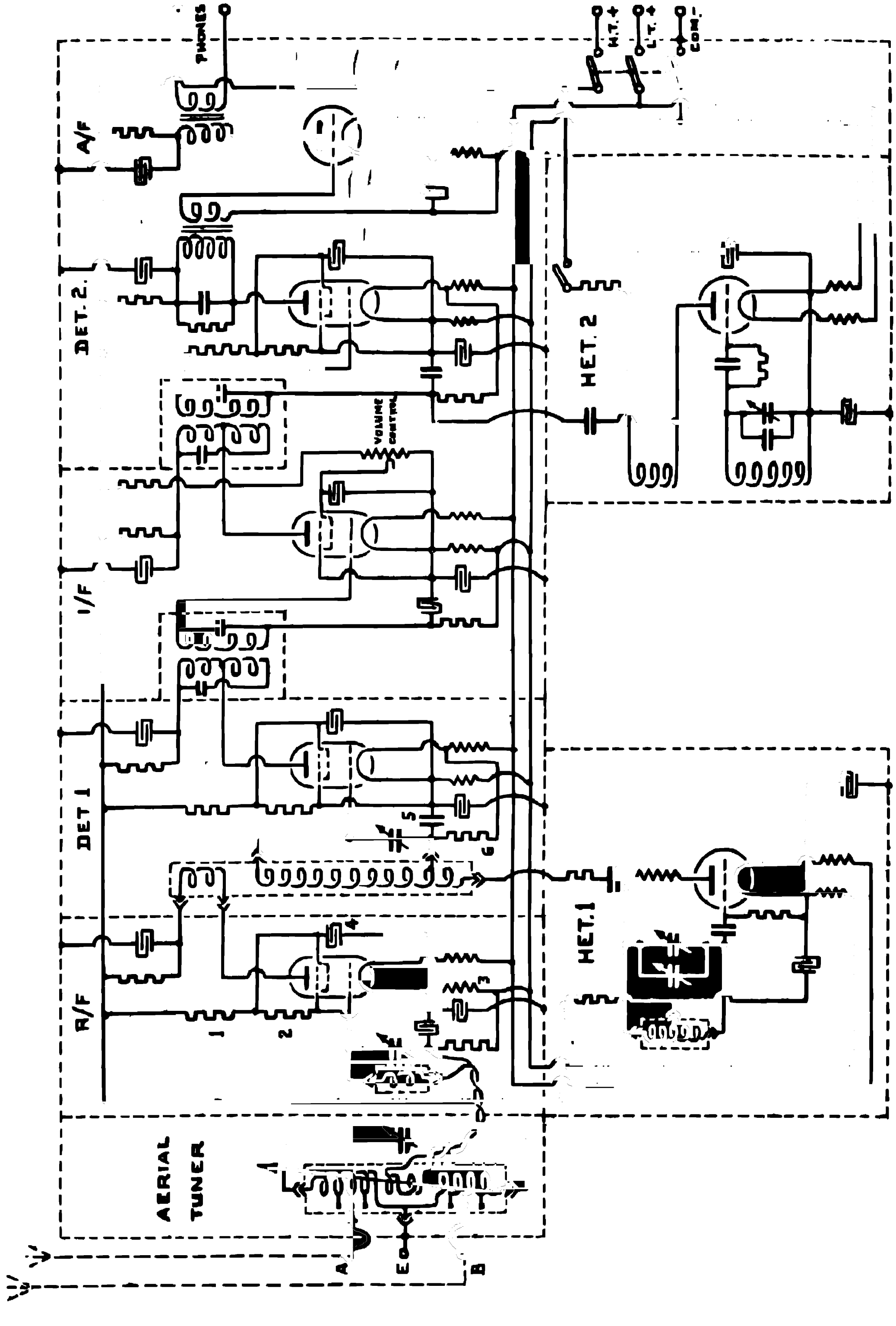

PHONES
H.T.+
L.T.+
COM.-
A/F
DET. 2.
HET. 2
VOLUME CONTROL
I/F
DET 1
HET.1
R/F
AERIAL TUNER
A
E
B
1
2
3
4
5
6

FIG. 38.

and earth. Intermediate values of the effective impedance can be obtained by varying the coupling between the aerial circuit and the R/F amplifier circuit. It is therefore possible to match approximately the impedance of any aerial system (*cf.* R.32), by choosing the appropriate tapping points and varying the coupling.

Where balanced aerials are not employed connection will usually be made to one of the aerial terminals, say A. The receiver may thus be used with open wire aerials of any length, with Beverage aerials, or distant aerials and arrays connected to the receiver through feeders, one side of which is earthed.

The R/F amplifier stage has a tuned grid circuit, tuned by a variable condenser with the value of 0·2 jar. The valve is a screened high frequency pentode, employing a 2-volt filament which is connected to a 4-volt supply through a 10-ohm resistance in each leg. These resistances serve to decouple the filament circuit of the valve, the negative side of which is shunted to the case through a by-pass condenser of value 0·1 μF. The potential drop in the resistance on the negative side also provides the requisite bias to the control grid of the valve, the connection being made through a 100,000 ohm resistance to the earthy end of the tuned circuit. Resistances (1) and (2), together with low resistance (3), form a potentiometer across the H.T. supply, and give the requisite potentials to the screen grid and suppressor grid respectively. The screen is connected to the filament negative, lead by the by-pass condenser (4).

The output of the R/F valve is applied to the grid circuit of the first detector valve by means of a transformer coupling.

The H.T. supply to the valve is decoupled by a 1,000-ohm resistance and a 0·1 μF condenser, an arrangement which is common to the other stages.

The first *detector* is also a *screened H/F pentode* valve, and the arrangement and decoupling of the supplies to the filament, screen, and anode, are similar to the R/F stage. The R/F transformer preceding this valve has a tuned grid inductance which has a third winding forming part of the coupling system between the R/F heterodyne and the first detector. The grid condenser (5) for the detector valve also forms part of the heterodyne coupling, and is connected between the earthy end of the grid circuit and the filament negative. The grid leak (6) has a resistance of 1 megohm, and is joined between the earthy end of the grid circuit and filament positive.

The R/F heterodyne has a Hartley circuit (K.7) employing a triode valve with a 2-volt filament, which is decoupled with two 5-ohm resistances in the legs, and a 0·1 μF condenser. The tuning condenser has a maximum capacity of 0·2 jar, but, for final tuning, a very small variable condenser is joined in parallel with the main tuning condenser. The anode of the valve is connected through a 0·0002 μF blocking condenser and a 5,000-ohm resistance to the coupling coil on the detector tuning inductance. This method of coupling is found to give sufficient heterodyne strength on all frequencies with the minimum of interaction between the heterodyne and signal frequency circuits. Since the heterodyne oscillator is separately controlled, it may be set either above or below the incoming signal frequency, if it is necessary to avoid second channel interference.

The output from the first detector valve is applied to the centre point of the primary inductance of the first S/F transformer. Both the primary and secondary are tuned to 120 kc/s. by semi-adjustable condensers. The coupling between primary and secondary is fixed, and gives approximately **optimum coupling** (paragraph 19) between the two tuned circuits under working conditions. The transformer is enclosed in a separate earthed copper screen. The secondary circuit of the first S/F transformer forms the grid circuit of the valve in the I/F stage. Negative bias is applied to the grid by the method used in the R/F stage. The screen grid supply is controlled by a 7,500-ohm variable potentiometer which is connected in series with a 20,000-ohm fixed resistance across the H.T. supply. This potentiometer is used to vary the amplification of the receiver (it alters g_m) and acts as a **volume control.** The output of the I/F valve is applied to the second detector by connection to a second S/F transformer similar in form to that already described.

The second *detector* is a *screened H/F pentode,* the input circuits to which are conventional in nature.

The output from the detector is applied to an output triode by means of a step-down transformer, the primary of which is tuned to approximately 1,200 cycles by means of a 1-jar condenser which also serves as an R/F by-pass. A 100,000-ohm resistance is also shunted across the transformer

primary circuit to flatten the tuning, which would otherwise be rather too sharp for aural reception of some H/F stations.

In the case of the triode output valve, the whole of the filament dropping resistance is placed in the negative leg in order to supply a greater negative bias to the grid. The A/F output is applied to 'phones through the usual telephone transformer.

The second heterodyne is required for C.W. reception, and it is generally advantageous to use it for I.C.W. also, since it greatly increases the efficiency of the detector on weak signals ; it should not be used when it is desired to maintain the quality of the I.C.W. note, or for listening to R/T transmissions. The self-oscillatory circuit is of conventional reverse feed type, employing a reaction coil and tuned grid circuit. The anode of the heterodyne valve is coupled through a 0·0002 μF blocking condenser to the grid condenser of the second detector valve. The frequency of the second heterodyne is set to 120 kc/s. ; a small variable condenser permits a change in frequency of about 5 kc/s. on either side, the frequency being 120 kc/s. when the condenser is in the middle of its travel.

In operating the receiver, the aerial coupling should first be set at its maximum value, and the four tuned circuits set to the desired frequency from the calibration curves. The second heterodyne should be switched on and its variable condenser set to the centre of its travel. The screen potentiometer (volume control) should be set to a position where valve noise is reasonably small.

With these settings, " searching " may be conducted by varying the setting of the condenser controlling the first heterodyne frequency. When the required signal is found, the four tuned circuits should be finally adjusted, and the aerial coupling reduced until the strength of signals just begins to fall. The fine adjusting condenser of the first heterodyne should then be used to bring the signal note to the dead space, after which the second heterodyne condenser may be detuned on either side to give a good readable note. Finally, the volume control may be adjusted to give reasonable signal strength. The band width passed by the intermediate frequency amplifier is of the order of 7 kc/s.

Second channel interference may occur from a powerful transmission differing by exactly 240 kc/s. from the signal frequency ; if present, it may be eliminated by setting the first heterodyne frequency on the opposite side of the signal frequency.

51. **Frequency Changing and Frequency Changer (Mixer) Valves.**—The frequency changer is the heart of the superheterodyne receiver and has been the subject of much development work since 1929 (B.43). The methods so far described are by no means the only ones available.

The general requirement of frequency changers may be put upon a broader basis by realising that, in all cases, **the object is to produce a difference or beat frequency between the signal frequency and that of the heterodyne oscillator**; in simple receivers without a separate oscillator frequency control, the oscillator frequency is usually the greater of the two. From the work in Section " D " on detection, it will be realised that beat frequency currents in the anode circuits of valves depend for their presence on the asymmetrical working of the latter ; the anode current waveform must resemble that shown in Fig. 36 to represent the mean anode current after the first stage of detection, suitably modified if anode bend detection is used.

For the purposes of classification, frequency changing may be accomplished in any of the following three ways :—

(A) The signal and heterodyne inputs may be injected in series between the cathode and control grid of a valve working as a detector, usually of the anode bend type. (**Grid injection**, and **cathode injection**).

(B) The signal input is applied between the control grid and cathode of a valve working as a detector, and the heterodyne oscillation is applied to the anode circuit in series with the H.T. voltage. (**Anode injection.**)

(C) The signal and heterodyne inputs are separately applied to two grids of a **multi-electrode valve** ; the latter may be of simple or complex construction. (**Electron coupling.**)

The method of frequency changing to be employed depends upon general design considerations, space, and cost. It may be said that, where the latter two conditions permit, the best results are usually obtained by using two separate valves, one to generate the oscillation and the other to "mix" the frequencies. The requirements of the modern domestic R/T superheterodyne receiver have been the main stimulus in the development of the **single valve frequency changer,** but it should not, however, be inferred that the latter possesses no advantages apart from a considerable economy in space and cost.

In all cases the oscillator must be as free from harmonics as possible, in order to avoid the introduction of **audible whistles** produced by unwanted transmissions. The latter difficulty is the source of one of **the faults** which form a criterion of comparison between different receivers; other faults include a tendency to **cross modulation** on strong signals, and the presence of **unwanted coupling** between the oscillator and signal circuits. The latter produces pulling or interlocking of these circuits when receiving weak stations, or working on high frequencies.

TYPICAL CASES OF (A).—Figs. 35 and 38 illustrate the series injection of the two inputs between cathode and control grid of the mixer valve. (**Grid injection.**)

Fig. 39 (*b*) represents a one-valve frequency changer, a tetrode, employing the coupling known as **cathode injection.** Considering the oscillator portion of the circuit, components (2), (3) and (4)

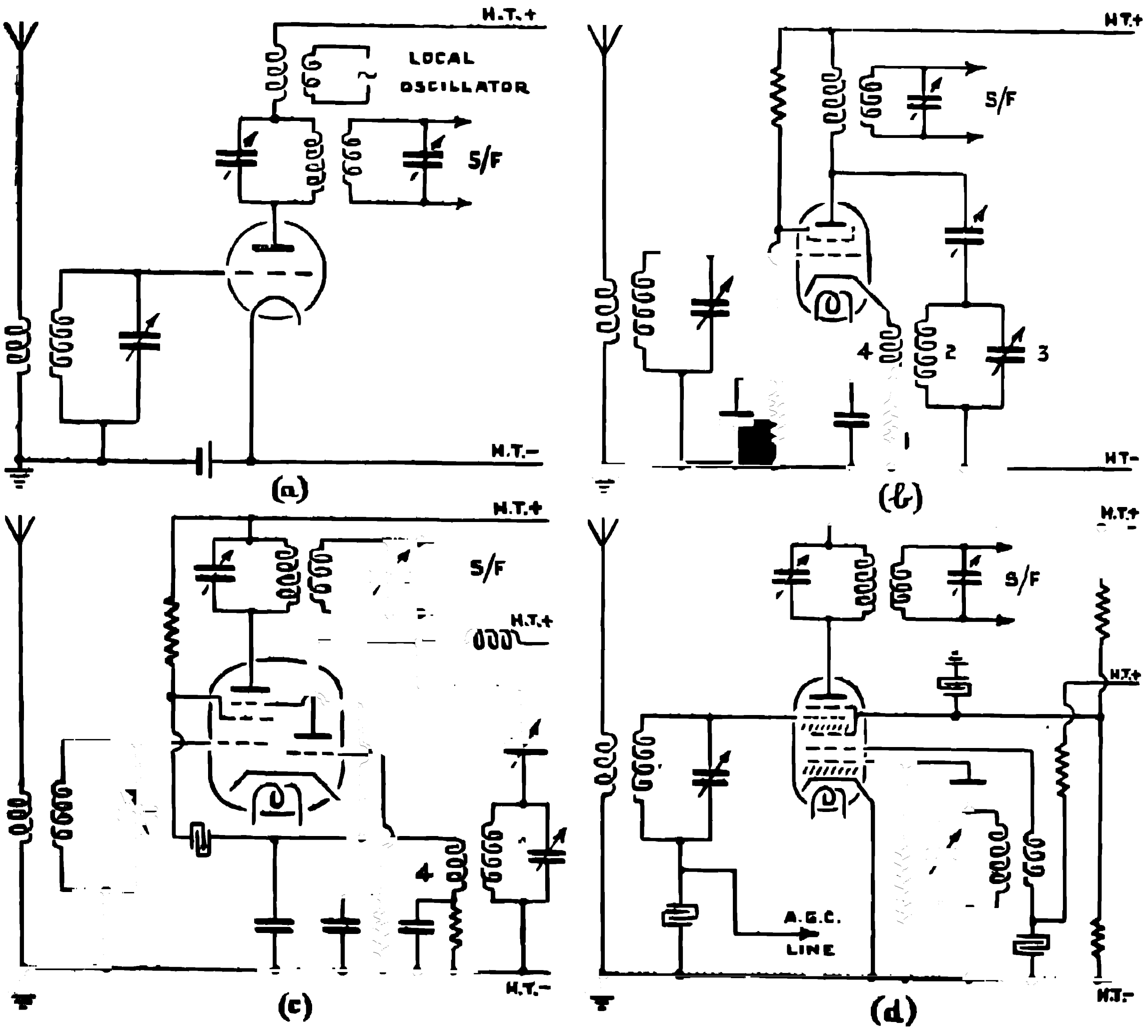


FIG. 39.

constitute a self-oscillatory circuit of ordinary reverse feed type, described as tuned circuit between anode and cathode with mutual inductive grid excitation (K.21). The cathode coil (4) is common both to the oscillator and signal input circuits ; automatic grid bias is provided by resistance (1), and the valve functions as an anode bend detector, giving rise to the usual high frequency terms in the anode circuit, together with sum and difference terms of the oscillator and signal inputs (Section "D," and N.19). The difference frequency component is selected by the tuned secondary circuit in the usual way.

Fig. 39 (*c*) represents a more complex method of achieving the same result. The valve is a **triode-pentode,** the triode portion being the oscillator, and the pentode portion the detector. Cathode injection is again employed, the cathode coil (4) being common to the oscillator and signal input circuits. This valve was developed in 1934, and may be considered to represent the best features of the single valve frequency changers employing an external coupling between the two portions. Anode bend detection of the two inputs in series provides the requisite beat frequency currents in the anode circuit.

It will be observed that there is essentially no difference between grid injection circuits (Fig. 35) and cathode injection. In the case of the latter, the small P.D. at oscillator frequency which is developed between the cathode and anode produces a result which is unimportant, since the mutual conductance is so much bigger than the anode conductance.

TYPICAL CASES OF (B).—Fig. 39 (*a*) represents what is known as **anode injection.** In this case only the signal frequency is rectified When the separate oscillator is not functioning, the receipt of a C.W. signal produces an R/F component in the anode circuit and gives a steady increase in the mean value of anode current, the value of which depends on the square of the input grid voltage, assuming the mean anode volts to remain constant. When the local oscillator is working, it injects an oscillatory component into the anode voltage which works into phase and out of phase, at the beat frequency, with the R/F component from the signal input. When they are in phase the mean anode voltage is greatest and the increase in mean anode current due to rectification of the signal is least ; the mean anode current therefore also varies at the beat frequency, due to a process which is stroboscopic in action, and to a valve which is adjusted to work asymmetrically, *i.e.*, as a detector. The circuit is one in which radiation from the local oscillator is minimised ; it is, however, seldom used. It may be noted that so long as the mixer valve is made to work asymmetrically, it does not matter which of the inputs is rectified, the signal input, the heterodyne, or both.

TYPICAL CASES OF (C).—Fig. 39 (*d*) represents a heptode (*cf.* N.63, Fig. 41), a frequency changer employing a somewhat different principle The disadvantages of previous frequency changers can mostly be traced to the nature of the coupling between the oscillator and signal frequency circuits, and to the necessity for detection at the signal input grid. In this case there is no common external coupling between the two circuits, the **coupling being internal and electronic in** nature ; moreover, detection at the signal grid may be avoided. The process is akin to modulation, especially **suppressor grid modulation** (N.23), or to the modulation method styled **screen grid injection.**

Electronic coupling was developed during 1932 and represented a great step in advance in frequency changer technique. The cathode with the first and second grids (B.43) forms the oscillator portion of the valve, the latter employing a reverse feed self-oscillatory circuit of conventional type. Electrons emitted from the cathode are controlled in their flow towards the oscillator anode by the oscillator grid. A portion of the electron stream passes to the oscillator anode, and the remainder passes on through the first screen grid to the signal grid (control grid), where it is further modulated by the input signal. After passing through the second screen, the electron stream arrives at the anode to which is connected the output impedance. It is sometimes considered that the valve has two space charges, shown shaded in the diagram ; there is the usual space charge in the neighbourhood of the cathode, and an oscillating supply of electrons constituting a "**virtual cathode**" in the neighbourhood of the control grid.

Electron coupling provides certain definite advantages ; circuit design is simplified, undesirable coupling between the two circuits is mainly eliminated, the oscillator is more stable, radiation of energy at the heterodyne frequency is decreased, and conversion conductance is higher.

A physical picture of the action of the heptode in producing ***beat frequency currents*** in the anode circuit may be obtained by consideration of the characteristic curves shown in Fig. 40 (*a*). These represent approximately the form of ordinary static characteristics plotted for variable signal grid (modulator grid) volts V_{g2}, and fixed values of oscillator grid volts V_{g1}. Let F represent the working point when no signal is applied either to the oscillator or signal input grids. With fixed signal grid volts, an oscillator input of 6 volts peak value would swing the anode current from G to E. A simultaneous input at the signal grid of 2 volts peak value would swing the anode current

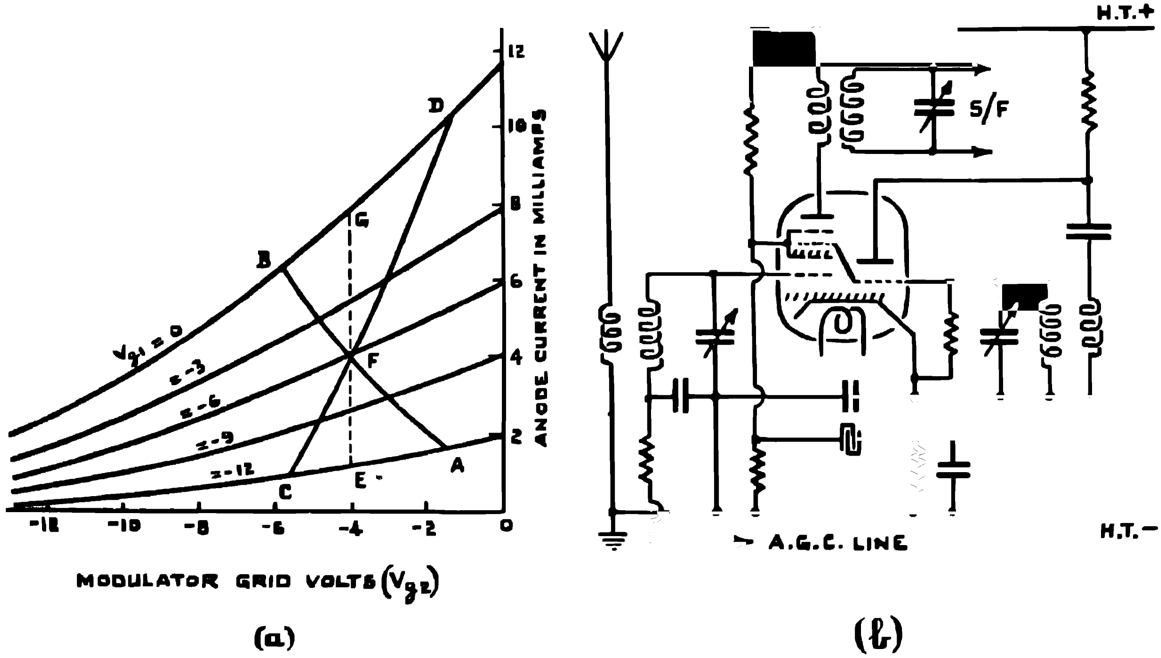


FIG. 40.

along the curve from B to A or from D to C, depending on whether the two inputs were in anti-phase or in-phase respectively. The oscillator and signal inputs may be considered to be the same in frequency but with a continuously varying phase displacement, the two inputs working into phase or out of phase with each other at a frequency equal to the difference of the two, the beat frequency. **The mean anode current corresponding to CD, a dual in-phase input, is greater than the mean anode current corresponding to the case of phase opposition, the curve BA.** The anode current therefore contains a beat frequency component which may be selected by a suitably tuned secondary circuit.

It may be noted that as the valve is made to traverse (say) the curve AB, the signal grid mutual conductance changes from a comparatively low value at A to a much higher value at B ; **the greater the input signal the greater will be the change in g_m, and the greater the amplification, the latter following almost a linear law.** Since, essentially, no detection takes place at the modulator grid, troubles due to cross modulation are largely overcome.

Fig. 40 (*b*) shows a **triode-hexode,** and represents the latest stage of development of one valve frequency changers. Electron coupling is again employed, and the arrangement of the electrodes is designed still further to minimise unwanted coupling between the two circuits.

52. **Principles of Super-regenerative Receivers.**—The use of ordinary regeneration or reaction as a method of amplification was shown in Section " D " to be equivalent to reducing the effective resistance of the input circuit. In the simple regenerative circuit it was proved that the

resistance was decreased by an amount $\frac{Mg_m'}{C}$, the effective resistance then being $R - \frac{Mg_m'}{C}$. When the reaction receiver is brought to the oscillating point, the effective resistance is zero, and the generation of self-oscillations is possible. With further increase of reaction, self-oscillations can build up to a larger amplitude, and while they are building up the effective resistance is negative, *i.e.*, $\frac{Mg_m'}{C} > R$. In all these cases, however, whatever may be its value, the effective resistance remains approximately constant during the operation of the receiver. The special feature of receivers using the principle of super-regeneration is that the effective resistance of the oscillatory circuit is deliberately varied at some chosen frequency, so that its value is alternately positive and negative, *i.e.*, free oscillations are periodically allowed to build up and to be damped out at a definite rate.

There are obviously two ways in which this may be accomplished :—

(1) The energy fed into the oscillatory circuit by reaction (the negative resistance) may be kept constant, and the positive resistance of the circuit varied at the chosen frequency so that it is alternately greater or less than the negative resistance.

The most elementary way of achieving this would be to have an additional resistance which could be inserted in, and cut out of, the oscillatory circuit at the frequency of variation desired, by some device such as a buzzer wheel. The constant resistance in the circuit would be less in value than the "negative" resistance, and the extra resistance would be sufficient to increase the total beyond the value of the "negative" resistance, so that the effective resistance would be alternately positive and negative.

Another possible method would be to connect across the tuned circuit a separate valve, the A C. resistance of which was varied at the desired frequency so as to constitute a variable damping device across the tuned circuit, and thus to cause it to alternate between a self-oscillatory and a non-oscillatory condition.

(2) The energy fed into the oscillatory circuit by reaction may be caused to vary periodically in value, so that it is alternately greater and less than the amount required to overcome the damping losses of the circuit, *i.e*, the negative resistance becomes alternately greater and less than the positive resistance. This is the method employed in Service receivers. The actual ways of accomplishing it will be dealt with after the nature of the results to be expected have been considered.

53. **Service Types of Receiver.**—Service super-regenerative receivers may be divided into two main categories :—

(*a*) Receivers in which the periodic variation of effective resistance is produced by means of a circuit other than that in which self-oscillations are alternately allowed to build up and die away. These are known as **quench receivers.**

(*b*) Receivers in which the effective resistance of the oscillatory circuit and its associated valve is arranged to be automatically self-adjusting, so that self-oscillations alternately build up and are damped out without the necessity of a separate circuit to vary the reaction. These are called **self-quenching receivers** or **squeggers.**

54. **Quench Receivers.**—In discussing the operation of a quench receiver it will be useful to recall to mind the nature of the current produced by the application of an alternating E.M.F. of amplitude $\mathcal{E}$ to an acceptor circuit (*i.e.*, an oscillatory circuit tuned to the applied E.M.F.), whose resistance is so small that the difference between its natural and resonant frequencies may be neglected.

The magnitude of the current at any time t after the E.M.F. is applied was found in Vol. I to be

$$i = \frac{\mathcal{E}}{R}\left(1 - \epsilon^{-\frac{R}{2L}t}\right) \sin \omega t.$$

The nature of the current is best grasped by considering the two terms of this result separately. It consists of

(1) A **forced** oscillation $\frac{\mathcal{E}}{R}\sin \omega t$, of constant amplitude if $\mathcal{E}$ and R are constant.

(2) A **free** oscillation of amplitude $\frac{\mathcal{E}}{R}e^{-\frac{R}{2L}t}$. The initial amplitude (at $t = 0$) of the free oscillation is $\frac{\mathcal{E}}{R}$, the same as that of the forced oscillation.

If the effective resistance of the circuit is positive, the free oscillation rapidly decreases in amplitude. This is the normal case in the receiving circuits so far considered, even in circuits with regenerative amplification nearly up to the oscillating point. The forced oscillation alone is then of importance, and regenerative amplification may be considered merely as a means of reducing the value of R so as to increase the amplitude $\frac{\mathcal{E}}{R}$ of this oscillation.

When the effective resistance is negative, however, the result is very different. The free oscillation then builds up as quickly as it was damped out in the case of positive resistance, and it is the forced oscillation which may soon be neglected by comparison.

This building up of free oscillations has already been considered at length in the section on the valve as generator. They would theoretically build up to an infinite amplitude, given an infinite time. In practice, the amplitude is limited in self-oscillatory valve circuits by the decrease of the negative resistance as the amplitude of oscillations increases, until eventually any further increase in amplitude makes the effective resistance again positive; oscillations are then maintained at a constant amplitude. But before these limiting conditions supervene, the following points are to be noted concerning the free oscillations :—

(*a*) They must be initiated by some momentary electrical disturbance of the oscillatory circuit, corresponding to the application of an E.M.F., as considered above. An incoming signal will perform this function, but the ordinary disturbances in a valve circuit will have a similar effect.

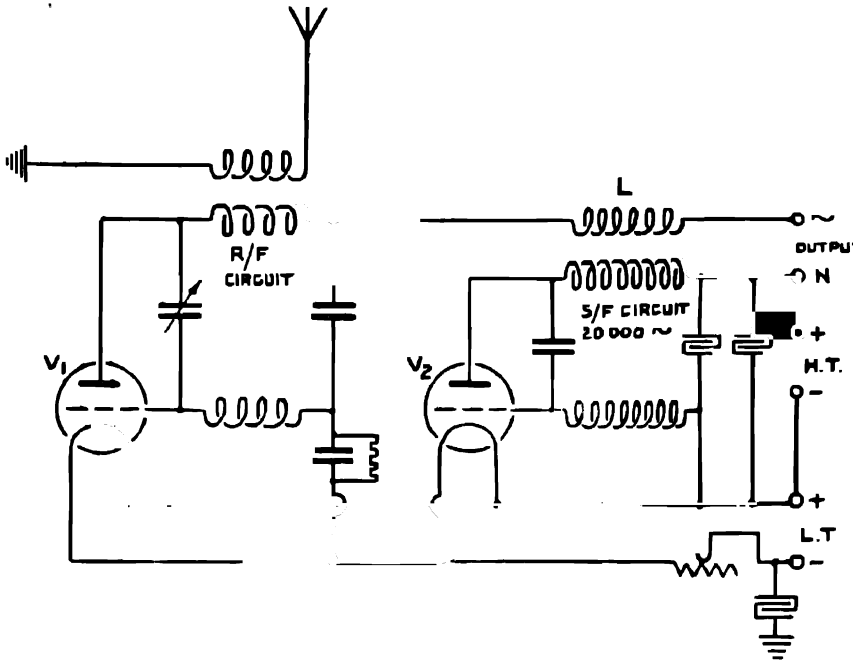
(*b*) Once they begin, they will build up as the resistance remains negative, whether the original stimulus is removed or not. If the latter is continued, as in the case of an incoming signal, the forced oscillation it produces is negligible in comparison.

(*c*) The initial amplitude of the free oscillation is the same as that of the forced oscillation $\left(\frac{\mathcal{E}}{R}\right)$, *i.e.*, proportional to the applied electrical disturbance, and while the oscillation is building up its amplitude retains this proportionality, for after any time t from the commencement of free oscillations it is $\frac{\mathcal{E}}{R}e^{-\frac{R}{2L}t}$.

The third conclusion above ceases to be true if the time t, during which the effective resistance of the circuit is negative, is long enough to allow the free oscillations to build up until they saturate the valve. Once this occurs, the free oscillations continue with a constant amplitude determined only by the conditions of the valve circuit (H.T. supply, saturation current, etc.), and bearing no relation to the amplitude of the initial disturbance or incoming signal.

Service quench receivers are only used for the reception of modulated signals, viz., I.C.W. signals, and so these will first be considered. A short account of the modifications necessary for the reception of C.W. will be given later.

55. **Reception of I.C.W. Signals.**—A simple quench receiver is shown in Fig. 41. The oscillatory circuit associated with the valve V_1 is of a nature capable of self-oscillation, the circuit being of the series feed, direct grid excitation type discussed in Section "K" (6, 7 and 8). It is tuned to the frequency of the incoming signal by the variable condenser shown. A separate oscillatory circuit of the same type is maintained in self-oscillation at a supersonic frequency



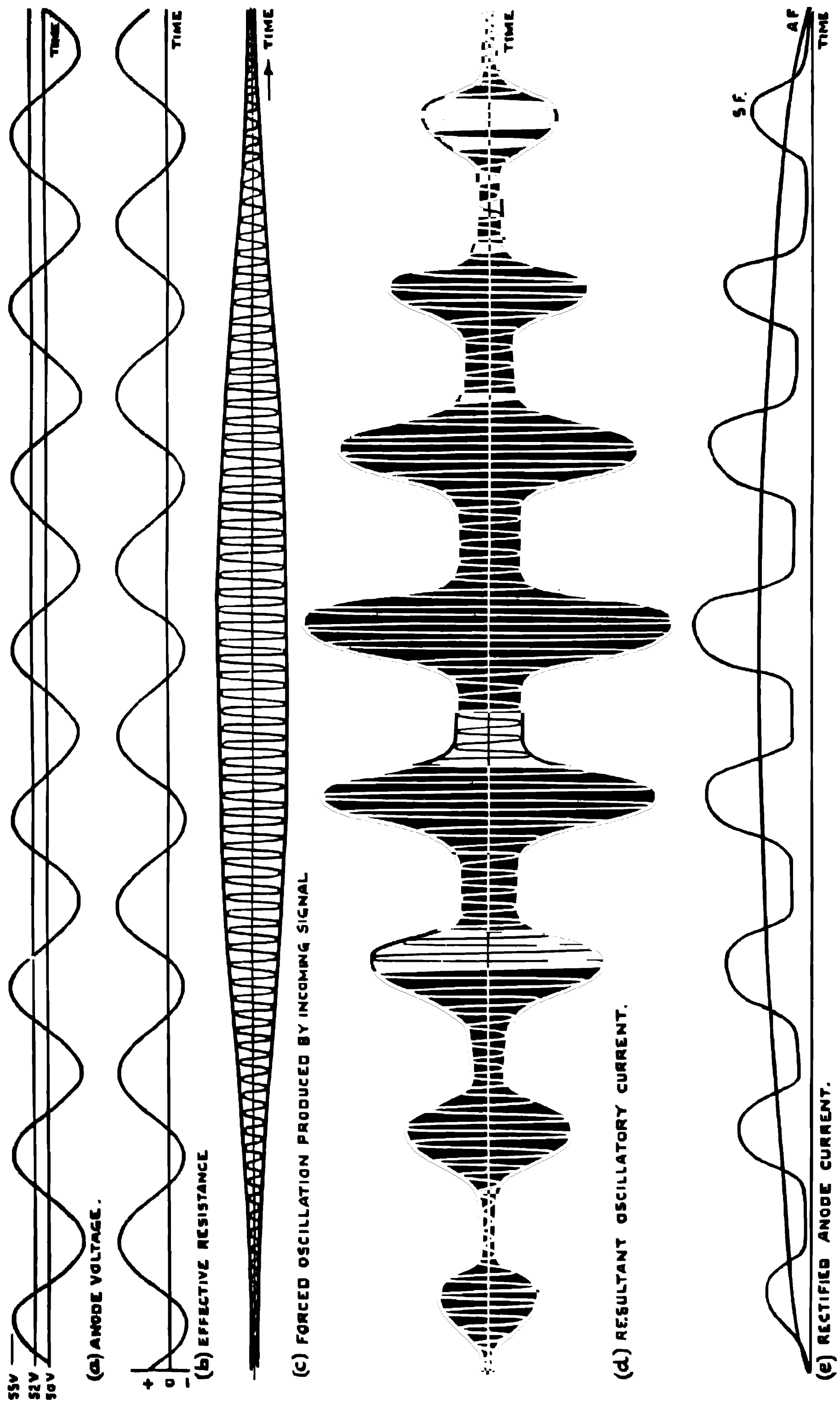
QUENCH RECEIVER.

FIG. 41.

(20 kc/s. in the circuit shown) by means of the valve V_2, and is mutually coupled to the coil L in the anode lead of the valve V_1. In consequence, the anode voltage of the valve V_1 is modulated at the frequency of the V_2 circuit, and alternates above and below the mean value of the H.T. voltage by the amplitude of the supersonic induced E.M.F. in the anode lead.

It will be remembered that although, for simplicity, we normally consider the characteristics to be parallel, the mutual conductance (g_m) of a triode actually increases as the anode voltage is increased, and it is on this fact that the possibility of varying the effective resistance of the V_1 circuit in this receiver depends. The slope of the dynamic characteristic g_m' is, of course, proportional to g_m, and therefore to the anode voltage. The "negative resistance" or energy fed into the oscillatory circuit is proportional to g_m', and so, ultimately, to the anode voltage. Thus the supersonic variation of anode voltage can be arranged to produce a supersonic variation of negative resistance above and below the positive resistance of the circuit, and therefore to make the effective resistance alternately positive and negative at the supersonic frequency. Approximate values in this receiver are as follows: the steady anode voltage is 50 volts; the 20 kc/s. induced E.M.F. in coil L has an amplitude of 5 volts. When the amplitude of anode voltage above is about 52 volts, free oscillations build up, *i.e.*, the nett resistance is negative; below this value of anode voltage, free oscillations are damped out. The variations of anode voltage and effective resistance are shown in Fig. 42 (*a*) and (*b*). Fig. 42 (*c*) shows the forced oscillation produced by the incoming I.C.W. signal voltage over one cycle of its audio frequency modulation. In Fig. 42 (*d*) is shown the alternate building up and damping out or "quenching" of the free oscillations. The points to notice are :—

(1) The final amplitude which the free oscillation reaches during any one period of negative resistance is proportional to the amplitude of the forced oscillation, *i.e.*, of the incoming signal at the commencement of that period. This depends on the fact that the free oscillation is quenched before its amplitude nears the condition when the valve is saturated.



QUENCH RECEPTION OF I.C.W. SIGNAL.
SONIC MODULATION

FIG. 42.

(2) Since the negative effective resistance is not constant over this period, but rises from zero to a maximum and falls again to zero, the free oscillations do not build up exactly according to an exponential law. Their rate of growth is largest when the effective resistance of the oscillatory circuit has its maximum negative value, and is lowest at the beginning and end of the self-oscillatory period, but the final amplitude of the free oscillation in every such period always bears the same ratio to the amplitude of the forced oscillation at the beginning of the period.

(3) The **free oscillation must be completely quenched during the period of positive resistance.** If there is a residual free oscillation when the effective resistance again becomes negative, the initial amplitude from which free oscillations build up is not that of the forced oscillation alone, but is the resultant of the forced oscillation and the remaining free oscillation. In consequence, the final amplitude of the free oscillation during a period of negative resistance is not proportional to the amplitude of the forced oscillation at its commencement, *i.e.*, is not proportional to the amplitude of the incoming signal at that instant. The mean audio frequency variation of the current in the oscillatory circuit, as shown in Fig. 42 (*e*), would then no longer necessarily bear any relationship to the modulation of the incoming signal.

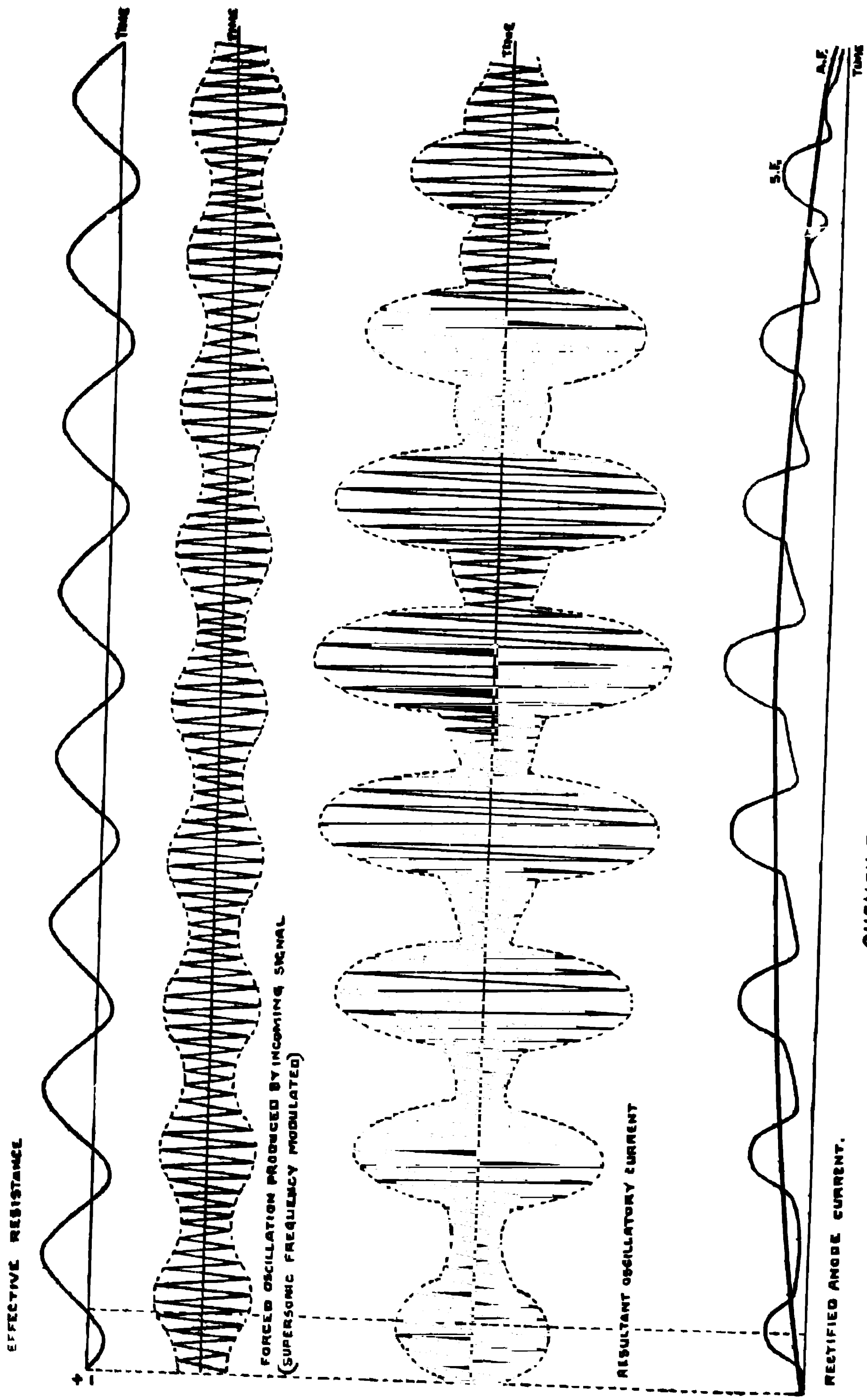
It should be noted that Fig. 42 (*e*) is diagrammatic only, since with cumulative grid detection the mean anode current should show a decrease.

(4) The measure of the amplification produced is obviously the ratio of the final amplitude of the free oscillation to that of the forced oscillation at the beginning of the period of effective negative resistance, and this can be made very large. It should be observed, however, that any atmospheric, interfering signal, or local disturbance such as that caused by irregular filament emission, will likewise initiate free oscillations if it occurs during such a period. The forced oscillation due to such disturbances (except atmospherics) will normally be of smaller amplitude than that due to the wanted signal, but the ratio of the two is preserved, and a background of " mush " will be heard in the telephones. The **selectivity** of a super-regenerative receiver is thus not outstanding. The weaker the signal the noisier is the background, if the signal is amplified to the same degree of audibility in each case.

56. **Choice of Quenching Frequency.**—The above considerations allow some simple deductions to be made as to the value of the quenching frequency, *i.e.*, the frequency with which the effective resistance of the receiving circuit varies between a positive and a negative value. The higher this frequency, the more often do oscillations build up and die away during one incoming wave train, and therefore the more faithful is the reproduction of the signal waveform. The limits of the ratio of quenching to incoming frequency are fixed by the amount of amplification required, the avoidance of saturation current, and the necessity that each set of free oscillations in the receiver should be damped out before the next set commences.

The higher the quenching frequency the less time there is for free oscillations to build up and die away, and so the smaller is the final amplitude reached in each burst of self-oscillation. *i.e.*, the less is the amplification. For instance, in receiving a wave modulated at 1,000 cycles per second, when the quenching frequency is 20 kc/s., self-oscillations build up and are damped out twenty times per audio frequency cycle. If the quenching frequency were doubled, the corresponding number would be forty times, *i.e.*, the time of building up self-oscillations and the time in which they are damped out would both be halved.

For a given quenching frequency the amplitude of the free oscillations increases with $\frac{R}{L}$, and likewise the rate at which they die away. The larger this ratio can be made, the higher can be the quenching frequency. To a first approximation $\frac{R}{L}$ increases with the radio frequency of the incoming



QUENCH RECEPTION OF I.C.W. SIGNAL
SUPERSONIC MODULATION

FIG. 43.

signal (to which the receiving circuit is tuned), and for a given signal frequency it increases with the stiffness of the circuit. Thus high amplification and faithful reproduction are more easily obtained the higher the incoming frequency, and the stiffer the tuned circuit. As the incoming radio frequency decreases, the amplification is reduced. Reduction of the quenching frequency may compensate for this up to a point, and at the expense of quality, but a limit is soon reached, for the quenching frequency must in practice remain supersonic. The amount of mush produced even at a supersonic quenching frequency has already been remarked, and when quenching takes place at an audible frequency, a continuous noise is produced in the telephones sufficient to drown any incoming signal. In addition, as the incoming radio frequency decreases, the selectivity of the receiver falls off considerably. The lowest frequency for which this method is of any value lies in the I/F band.

Various other methods of applying the super-regenerative principle to the reception to I.C.W. have been devised ; one method, for instance, employs a **supersonic modulation of the carrier wave** at a frequency which gives an A/F beat with the quenching frequency. For example, a carrier wave modulated at a frequency of 21 kc/s. would give an A/F beat note of 1 kc/s. if the quenching frequency were 20 kc/s. ; the mechanism of this process is made clear by means of the curves of Fig. 43.

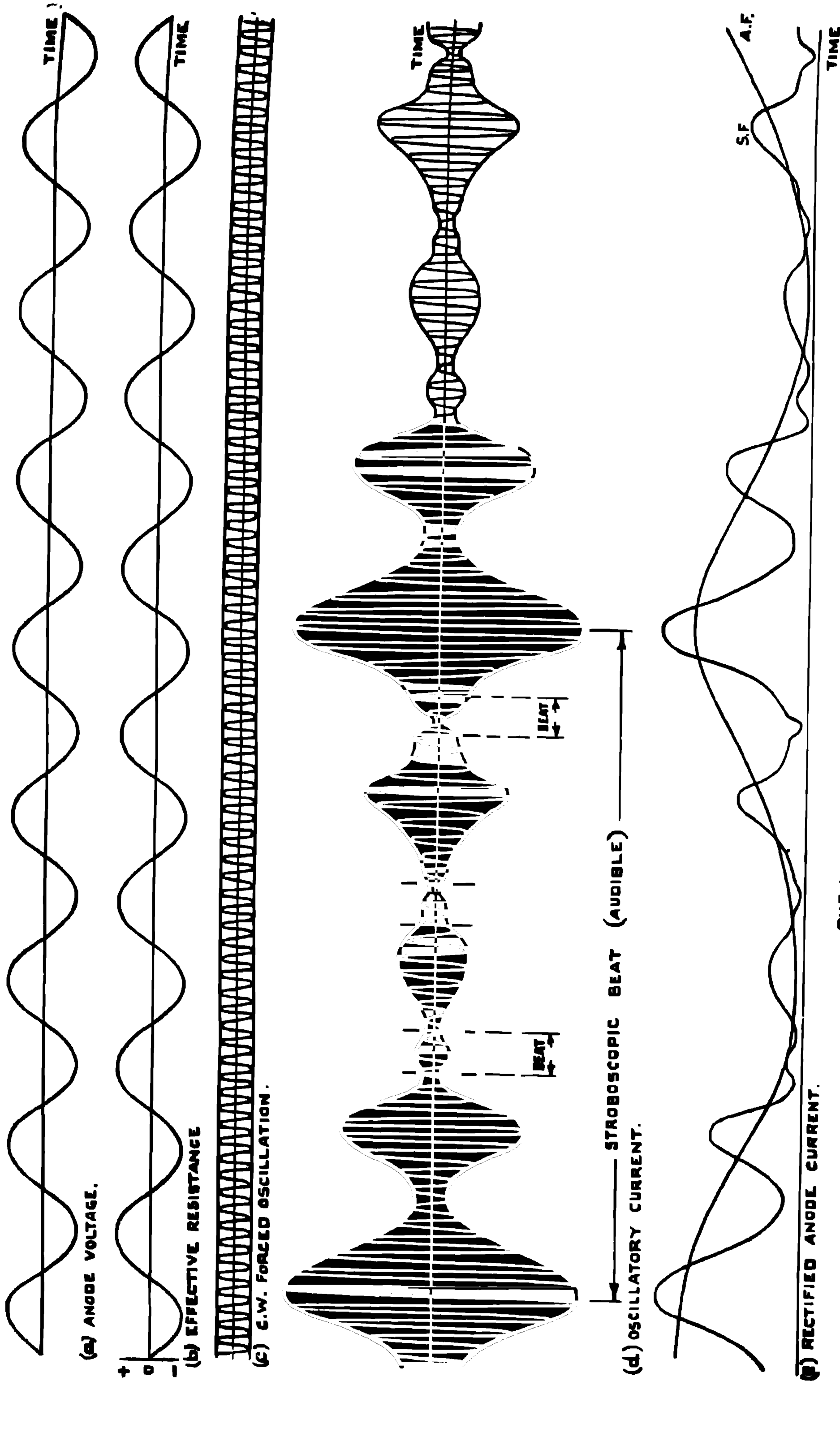
57. QUENCH RECEPTION OF C.W. SIGNALS.—The ordinary method of introducing an audio frequency variation of amplitude into a C.W. signal is to employ a separate heterodyne or an autodyne circuit. It is possible to use the quench receiver in this way, provided that the free oscillation is not completely quenched during the period of positive resistance, but still retains an amplitude comparable with that of the forced oscillation every time the resistance becomes negative. Both the free and the forced oscillations thus continue for the whole duration of the forced oscillation, and the maximum amplitude attained by the free oscillations, during a quenching cycle, is proportional to the sum of the amplitudes of the free and forced oscillations at the beginning of the negative resistance period. Provided that the free and forced oscillations are of the same order of magnitude at such instants, their relative phase then is the main factor in determining the amplitude to which free oscillations build up. For this to be true it is also necessary that the free oscillations should not reach a value which saturates the valve, just as in the I.C.W. case. The free and forced oscillations thus give a maximum resultant amplitude when they are in phase, *i.e.*, at the instants where beats occur ; at the beginning of intermediate quenching cycles they are not in phase, and the smaller resultant amplitude at these instants gives a smaller final amplitude of free oscillation. Hence the envelope of free oscillations over a number of quenching cycles is modulated at the beat frequency of the free and forced oscillation, and an audible note is heard when detection takes place.

The chief difficulty of the ordinary heterodyne or autodyne method of reception, at high radio frequencies, is that a small percentage change in the frequency of the incoming signal gives a large actual change in frequency, *e.g.*, at 10 Mc/s. a change of 0·1 per cent. corresponds to 10,000 cycles per second. Even if the frequency of the local oscillator remained constant (which is unlikely, especially in an autodyne circuit), the beat note would change by 10,000 cycles per second. It is easily seen that under practical conditions the beat frequency often becomes supersonic, and therefore inaudible, and continual searching is necessary to keep the signal audible. The peculiar feature of the quench receiver, when adjusted so that the free oscillation is never completely quenched, is that it is still possible to get an audible note on detection, although the beat frequency between the incoming signal (the forced oscillation) and the local free oscillation is itself well above the audible range. Thus sudden changes in the signal or local frequency do not involve loss of the signal, but merely a change in the pitch of the note heard in the telephones.

This phenomenon may be explained as follows. Suppose that f_1 is the incoming signal frequency and f_2 is the frequency of the free oscillation in the quench receiver. The beat frequency is then $f_1 - f_2$, and, as pointed out above, this is often inaudible. If ϕ is the quenching frequency, the time of one quenching cycle is $\frac{1}{\phi}$ and the number of beats per quenching cycle is therefore $(f_1 - f_2) \times \frac{1}{\phi} = \frac{f_1 - f_2}{\phi}$.

If this is a whole number, the phase difference between the dying free oscillation and the forced oscillation, at the beginning of a period of negative resistance, is always the same ; for instance, if they were in phase at the beginning of one such period, they would again be in phase at the beginning of the next period, and so on. After a few quenching cycles the initial amplitude from which free oscillations build up would be such that their final amplitude would saturate the valve. Each succeeding burst of self-oscillation would similarly saturate the valve as long as the forced oscillation persisted, and, since the quenching frequency is supersonic, no audible result would be produced.

The possibility of $\frac{f_1 - f_2}{\phi}$ being a whole number, however, is remote, as will easily be recognised, and so the phase difference between the free and forced oscillations is different at the beginning of consecutive period of negative resistance. Hence the amplitude reached by the free oscillations during consecutive quenching cycles



QUENCH RECEPTION OF C.W. SIGNAL.

FIG 44.

differs correspondingly; it is a maximum when the free and forced oscillations are in phase at the beginning of a period of negative resistance, and a minimum when they are in anti-phase at such an instant.

Nevertheless, over a number of quenching cycles, say n, the number of beats is $n\frac{(f_1 - f_2)}{\phi}$, and for some value of n this is bound to be a whole number. If, for simplicity, we assume that the free and forced oscillations are in phase at the beginning of the forced oscillation, they will again be in phase at the end of the "n"*th* quenching cycle. At the beginning of intermediate quenching cycles between these limits they are not in phase, and so the final amplitude of the free oscillations during these cycles is smaller. Hence the amplitude of the free oscillations reaches a maximum value every n quenching cycles, *i.e.*, it is modulated at a frequency equal to $\frac{\phi}{n}$. This frequency will in general be audible, and so an audible note is heard in the telephones on detection. Alteration in f_1 or f_2 will alter the value of n which makes $n\frac{(f_1 - f_2)}{\phi}$ a whole number, and so will alter the pitch of the note heard, but it will remain audible over a large range of values of f_1 and f_2. The process is illustrated for constant values of f_1 and f_2 in Fig. 44.

This kind of super-regeneration is sometimes said to be "stroboscopic." The actual phenomenon is probably more complicated than is indicated by the simplified explanation given above, but it remains of the same general nature. The essential conditions for its production are that the valve should not be saturated by the final amplitude of the free oscillations, and that these should only be damped to an amplitude of the same order as that of the incoming signal during the quenching period. This may be attained in practice by increasing the part of the quenching cycle during which the resistance is negative, and decreasing, if necessary to prevent saturation, the amplitude of variation of effective resistance about its zero value.

The advantage of this method of reception is that, even with large frequency variations in transmitter and receiver, the signal will still be heard. This involves, of course, that any interfering signals in the same range of frequencies are received with nearly equal intensity, and so the method is extremely unselective.

58. **Self-quenching Receiver.**—In this receiver self-oscillations are built up and quenched periodically, as in the quench receiver, but no separate self-oscillatory circuit is used to produce the necessary variation of negative resistance above and below the positive resistance of the receiving circuit. This is accomplished automatically by appropriate coupling between the anode and grid circuits and the use of a grid insulating condenser and leak.

The Oscillator.—The circuit of a self-quenching receiver is the same as that already discussed under regenerative amplification and autodyne reception (Section "D"), and is reproduced in Fig. 45. It will be remembered that with loose reaction coupling, amplification is produced, and, as the coupling is made tighter, a point is reached at which self-oscillations are generated. The building up of continuous self-oscillations, and the amplitude they acquire in a circuit with negative grid bias, is discussed in Section "K." In the oscillator then considered, the tuned circuit was between anode and filament, but the principle is the same when the tuned circuit is between grid and filament. It was shown that under favourable conditions the positive half-cycle of grid oscillatory voltage covered the whole straight part of the dynamic characteristic, and that the mean anode current then flowing was much greater than that corresponding to the same grid bias under non-oscillatory conditions. The maintenance of continuous oscillations depends on keeping the grid bias steady at the appropriate value.

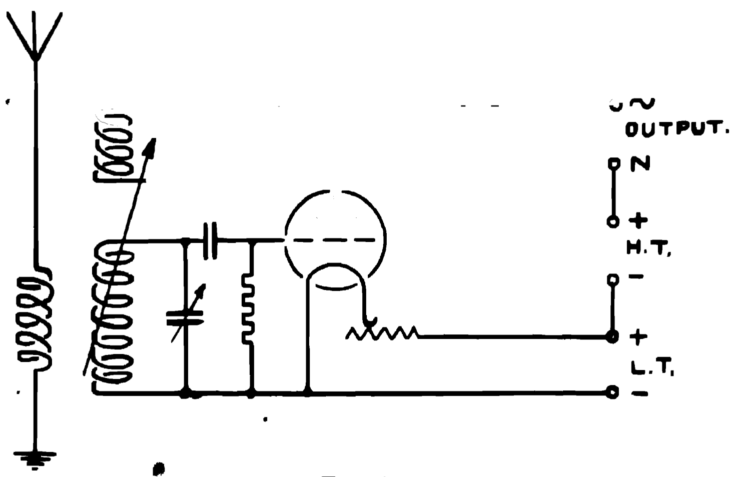
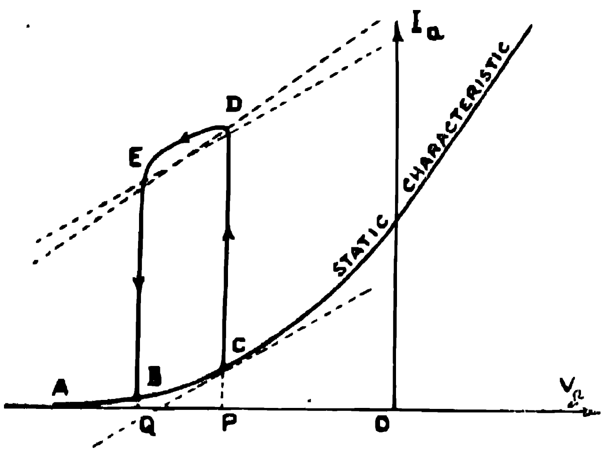


FIG. 45.

Suppose now that the coupling between anode and grid circuits is made still tighter. The immediate effect is to increase the amplitude of oscillatory grid voltage. This, however, causes an increase in the grid current flowing during the positive half-cycle; the steady potential of the grid therefore becomes more negative, and the positive peak of oscillatory grid voltage returns to approximately the same potential as before. The negative half-cycle of grid swing then extends further into the region where no anode current flows, and the result is that the slope of the effective dynamic characteristic, *i.e.*, the ratio of the total change in anode current to the total change in grid voltage, is decreased, the nett resistance of the circuit becomes positive, and oscillations are damped out. No grid current then flows, and, as the grid condenser charge leaks away through the resistance, the grid rises to a less negative potential and oscillations can be built up again. Thus, when the coupling between anode and grid circuits is made sufficiently tight, a circuit of this type automatically allows self-oscillations to build up and die away; the whole is said to exhibit the phenomenon of "**grid tick,**" or "**squegging.**" It will be seen that the kind of oscillation produced is that described as I.C.W., and if the self-quenching frequency is an audible one, the circuit may be used as an **I.C.W. transmitter,** the wave radiated from which will give an audible note in a receiver, without the necessity for heterodyne reception. When used in this manner, the tuned circuit is generally between anode and filament, as in the ordinary C.W. transmitter.

The Receiver.—The use of the circuit as a receiver will now be considered. In this case the tuned circuit is between grid and filament, as shown in Fig. 45. The possibility of reception depends on the fact that **the negative grid bias at which self-oscillations start to build up is appreciably less than that at which they are damped out or "quenched,"** *e.g.*, the values found in an experimental determination were V = — 14 volts for the quenching, and V = — 3 volts for the building up, of self-oscillations. The reason for this appears to be as follows. When oscillations are just beginning to build up, their amplitude is, of course, extremely minute. The oscillatory P.D. in the external anode circuit is correspondingly minute, and the anode voltage remains very nearly the same as under static conditions; thus the slope of the dynamic characteristic is practically the same as that of the static characteristic, and as the actual grid bias gives a working point beyond the lower bend of the characteristic, this slope is very small.



----- DYNAMIC CHARACTERISTICS.

FIG. 46.

As the oscillations build up, the oscillatory grid voltage extends into the straight part of the characteristic, and the mean anode current jumps to the much higher value associated with self-oscillations of large amplitude. For a steady grid bias of this order, the conditions are shown in Fig. 46. CP is the steady anode current for negative grid bias OP when oscillations are just beginning. DP is the mean anode current for oscillations of large amplitude. The mean dynamic characteristic corresponding to DP is shown by the dotted line through D (it is actually a loop). Its slope is greater than that of the static characteristic at C, which is approximately the g_m at the beginning of self-oscillations. As the negative grid bias is increased, the mean anode current decreases along DE, and oscillations decrease in amplitude. In the neighbourhood of E, the oscillations fall to such an amplitude that the slope of the dynamic characteristic through E is not sufficient to make $\frac{Mg_m'}{C} > R$. Oscillations are damped out, and the mean anode current falls from EQ to the much smaller value BQ, which corresponds to static conditions with a negative grid bias OQ. Thus if

the negative grid bias is greater than OP, oscillations will not build up, but if they have built up they will be maintained until the grid bias reaches the larger negative value OQ. (A mathematical explanation of this phenomenon has been obtained by representing the valve characteristic by an equation of the fifth degree.)

When the grid bias is made self-adjusting, by an insulating condenser and leak, the actual cycle of mean anode current values obtained is not CDEBC, for the jump of mean anode current from C to D is associated with the largest oscillatory grid voltages, and these cause grid current to flow and increase the negative grid bias. The curve connecting mean grid bias and mean anode current over one self-quenching cycle is thus as illustrated in Fig. 47 by the loop CGHKB. The mean grid potential becomes more negative during the part CGHK of the loop. After K, oscillations have almost died out, and the loss of charge through the leak resistance causes the grid bias to decrease, as shown by KBC. At C, oscillations again start to build up, and the cycle is repeated. The part CGH of the loop is traversed much more quickly than the remainder, and the part KBC, when the condenser is discharging, occupies much the greatest proportion of the time taken for one self-quenching cycle.

The initial process on making the H.T. supply is also shown in Fig. 47. The grid bias is then zero, and oscillations build up more rapidly, and rise to a higher amplitude, than when steady

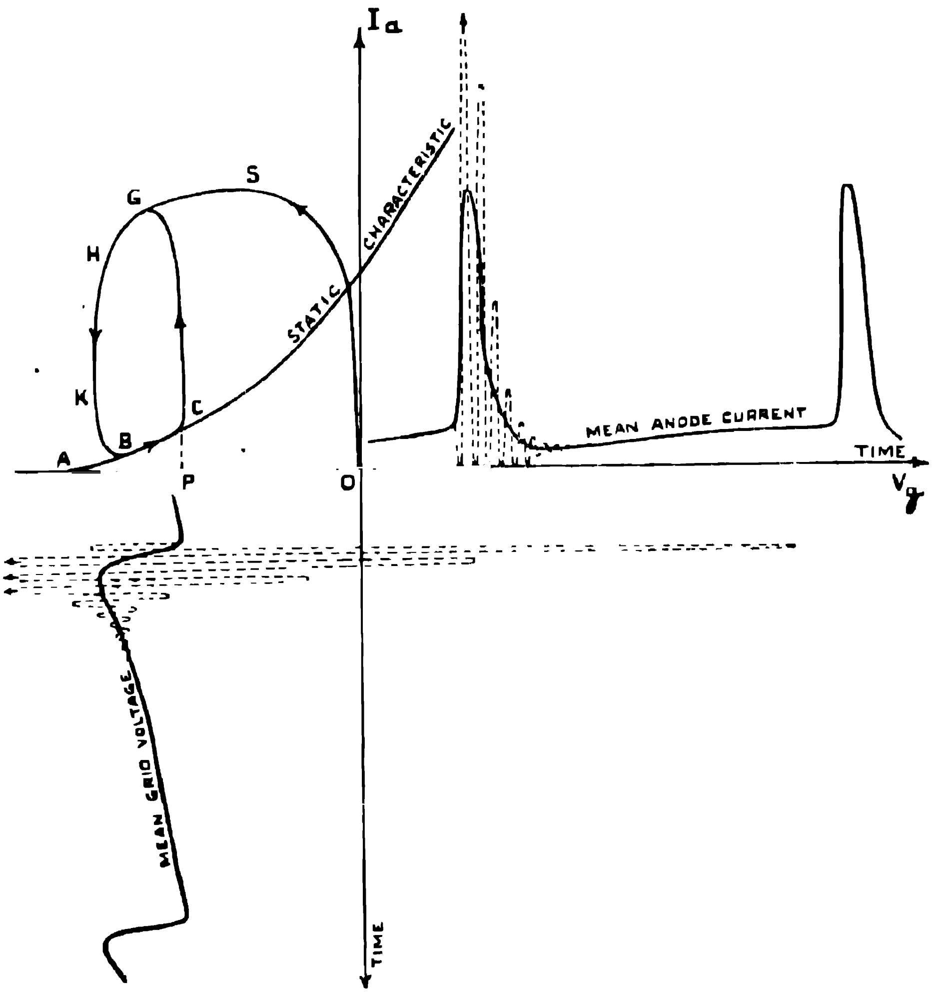


FIG. 47.

conditions have been reached. The curve connecting mean anode current and mean grid bias during this preliminary cycle is indicated by OSGHKBC.

An attempt has also been made in this figure to illustrate the actual variation of anode current and grid voltage with time over one self-quenching cycle. There is, of course, a very much larger number of R/F oscillations than it is possible to show in the figure.

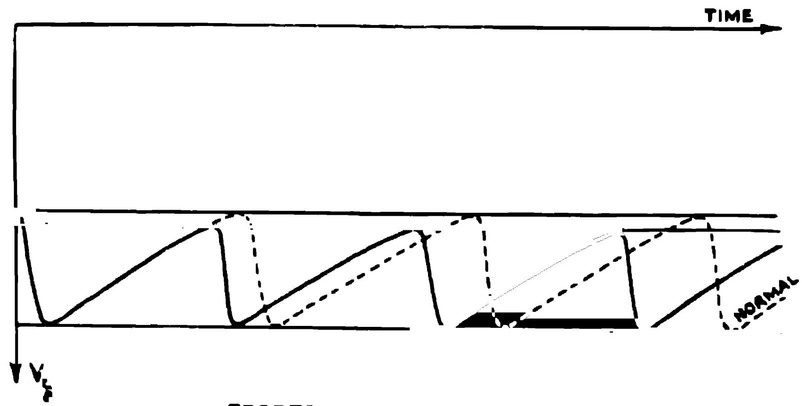
The time taken to traverse one self-quenching cycle depends mainly on the values of the grid insulating condenser and leak, being proportional to the time constant "CR" of this combination (Vol. I). It also depends slightly on the value of the reaction coupling, filament emission and anode voltage. By a suitable choice of C and R the self-quenching frequency may be made either audible or supersonic as desired. Alteration of either C or R is equally effective in varying this frequency. The frequency of the R/F oscillations is, of course, determined chiefly by the values of the inductance and condenser in the grid tuned circuit, as in the case of continuous oscillations.

Thus, if "CR" is such as to make the self-quenching frequency audible, a continuous note will be heard in the telephones ; but if the self-quenching frequency is supersonic, no such note will be heard. In both cases, however, there is always the strong back-ground of mush that has already been referred to under Quench Reception.

59. **Effect of an Incoming Signal.**—An incoming signal produces a forced oscillation in the grid tuned circuit. The effect of this is negligible while the circuit is in its self-oscillatory condition, just as in quench reception ; the quenching of free oscillations occurs at sensibly the same mean grid voltage, whether the forced oscillation is present or not. The forced oscillation, however, produces an appreciable change in the mean grid voltage at which free oscillations begin to build up, *i.e.*, the voltage OP in Fig. 47. As the condenser is discharging along BC, an oscillatory voltage of finite amplitude, due to the forced oscillation, is superimposed on the mean grid voltage, with the result that the slope of the dynamic characteristic increases sufficiently to make the effective resistance negative before the mean grid bias reaches OP. As this part of the self-quenching cycle takes place relatively slowly, the time between the instant when free oscillations are quenched and the instant when they start again is reduced by an appreciable amount. In other words, the period of the self-quenching cycles is decreased, and so their frequency is increased. The phenomenon has been compared with that of starting and sliding friction ; the frictional force opposing the movement of a block of (say) wood over a rough surface is greater when the block is at rest than when it is moving. If the self-quenching frequency is audible before the incoming signal arrives, the effect of this increase in frequency is to raise the pitch of the notes heard in the telephones (*cf.* W.15) ; if the self-quenching frequency is originally supersonic it will rise to a higher supersonic frequency.

It may be noted that a decrease of filament current also increases the frequency of the signal note.

60. **Reception of C.W. Signals.**—The above effect is sketched for a C.W. signal in Fig. 48, where only the mean grid potential variation with time is shown. The difference in grid bias for



RECEPTION OF C.W. SIGNAL

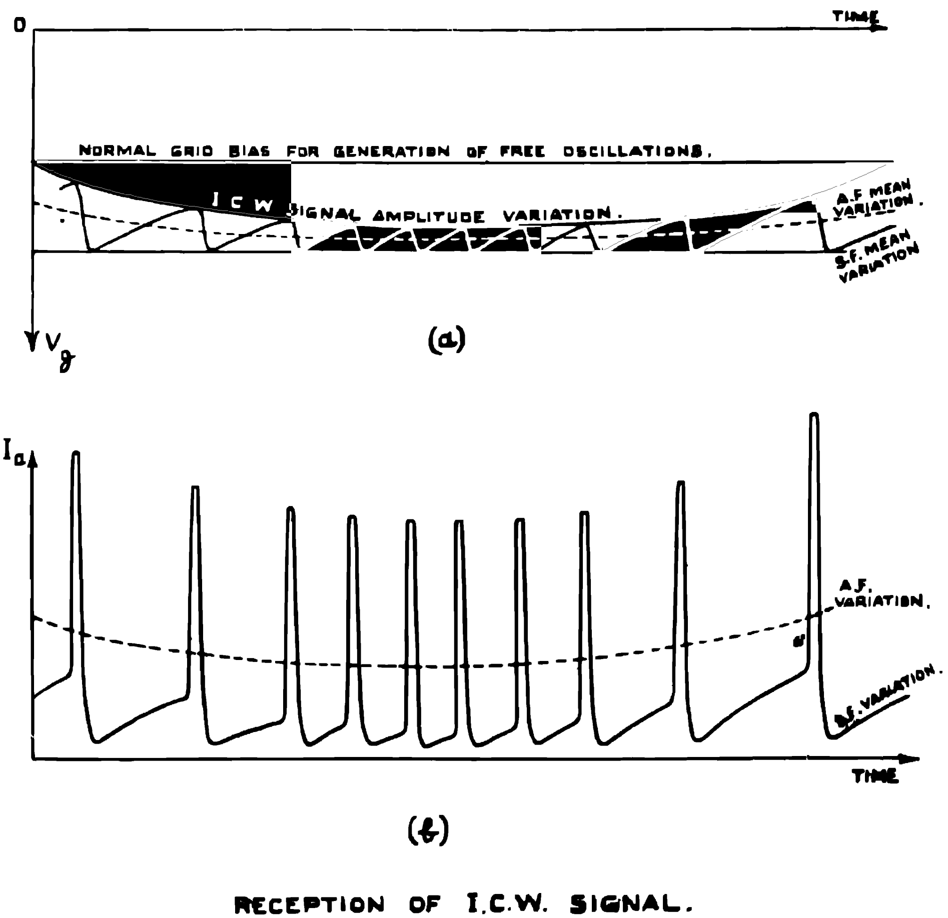
FIG. 48.

the generation of free oscillations, with and without a forced oscillation, may be considered roughly proportional to the amplitude of the forced oscillation. For a C.W. signal this is constant, and so the effect is as indicated (*cf.* W.15).

In order to receive C.W. signals, the normal self-quenching frequency of the receiver must lie in the audible range. A note of definite pitch is then continuously audible in the telephones. When the incoming C.W. signal arrives, the pitch of the note rises and the note remains at a higher pitch for the duration of the signal, which can therefore be read. This type of self-quenching receiver is also called the " **howling squegger.**"

★61. **Reception of I.C.W. Signals.**—In this case, the incoming signal possesses an audio frequency variation in amplitude, and so an audio frequency self-quenching cycle is not desirable. The I.C.W. waveform can be reproduced by the envelope of free oscillations if the self-quenching frequency is supersonic. One I.C.W. cycle, at say, 1,000 cycles per second, covers the same period of time, for instance, as twenty self-quenching cycles at a quenching frequency of 20 kc/s , and during these twenty cycles the amplitude of the forced oscillation (the I.C.W. signal), at the beginning of each self-quenching cycle, rises from a minimum to a maximum and falls to a minimum again.

The decrease in the period of a self-quenching cycle is proportional to the amplitude of the forced oscillation at the beginning of that cycle. With an audible quench frequency considerably higher than the I.C.W. modulation, this would be made evident by the pitch of the note heard in the telephones rising to a maximum when the I.C.W. amplitude was a maximum, and falling at the end of the I.C.W. cycle to its value at the beginning. With a supersonic quench the effect is of the same kind ; self-quenching cycles are completed most rapidly in the middle of the I.C.W. cycle, and so there are most positive pulses of anode current then. The result is that the mean anode current over a few quenching cycles is greater in the middle of the I.C.W. cycle than it is at the beginning and end, *i.e.*, there is an **audio-frequency increase** of mean anode current due to this cause.



RECEPTION OF I.C.W. SIGNAL.

FIG. 49.

Another effect of the I.C.W. signal amplitude variation, however, is that the grid bias at which oscillations start is most negative in the middle of the I.C.W. cycle. Over such a cycle the mean grid potential therefore has an audio-frequency variation which is in the negative direction, and so gives an **audio-frequency decrease** of mean anode current.

These two opposed effects occur simultaneously and the nett audio-frequency variation of anode current, which is the measure of the amplification produced, is proportional to their difference. By a suitable choice of circuit constants it is possible to cause the one effect to outweigh the other sufficiently to give efficient amplification.

Fig. 49 is an attempt to illustrate the reception of I.C.W. signals on a supersonic self-quenching receiver. Fig. 49 (*a*) shows the S/F cycles of grid voltage for part of an audio-frequency cycle of the forced oscillation. The envelope of the latter is greatly exaggerated in order to show the effect. It should, of course, be of much smaller amplitude than the mean audio-frequency variation of grid voltage produced by the generation and quenching of free oscillations, and shown by the dashed line. In Fig. 49 (*b*) are sketched the corresponding supersonic and audio-frequency variations of anode current. The effect of the increasing negative grid bias is taken to be greater than that of the higher frequency of positive anode current pulses, and so the audio-frequency variation is shown as a decrease.

62. High Speed Reception.—High speed reception of telegraphic signals may be accomplished by attaching to a receiver a D.C. amplifier and an automatic tape recorder called an *undulator*.

One of the principal difficulties consists in the fact that the undulator is a D.C. operated mechanism. In principle, its basic electrical feature is a moving coil loudspeaker, but in place of the diaphragm it has a pivoted lever coupled to the stylus or siphon arm. The moving unit operates in the vertical direction only, and carries two independent windings, or coils; these may be connected in series or parallel, for single current working using a spring bias, or they may be used separately, one for "**mark**" and the other for "**space.**" The latter represents the Service usage, the marking and spacing D.C. currents being arranged to operate the moving coil unit in opposite directions.

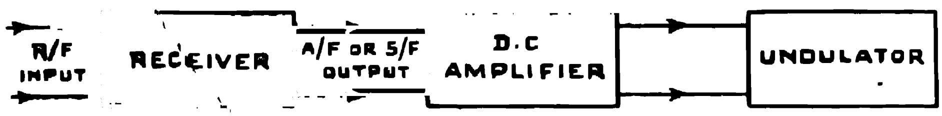


FIG. 50.

The A/F (or S/F) signal output of a receiver is, therefore, not directly available for application to the undulator; it must first be rectified, the small D.C. component of the anode current resulting from rectification being subsequently amplified by a **D.C. amplifier** (K.59) before application to the **undulator**. Fig. 50 represents a schematic diagram of a high speed receiver.

No extended account is given of the undulator, which is an instrument in common use in telegraphy. The unit of most importance and electrical interest is the D.C. amplifier, an account of its general principles being given below.

Fig. 51 represents the electrical details of a D.C. amplifier which may be called upon to perform the following four functions :—

(*a*) To rectify the A/F or S/F signals from the receiver, converting them to D.C. morse signals.

(*b*) To limit the amplitude of the D.C. signals to a suitable value independent of the amplitude of incoming signals. **The limiter** acts as a delayed auto-gain control, which comes into operation when input signals exceed a certain value, but provides a timed delay in releasing the control at the end of a signal. This permits fading signals to be recorded without distortion.

(*c*) To amplify the D.C. output from the rectifier and convert it to spacing and marking currents for double current working of the undulator.

(*d*) To provide a shaping control of the dots, spaces and dashes which may be necessary at high speeds.

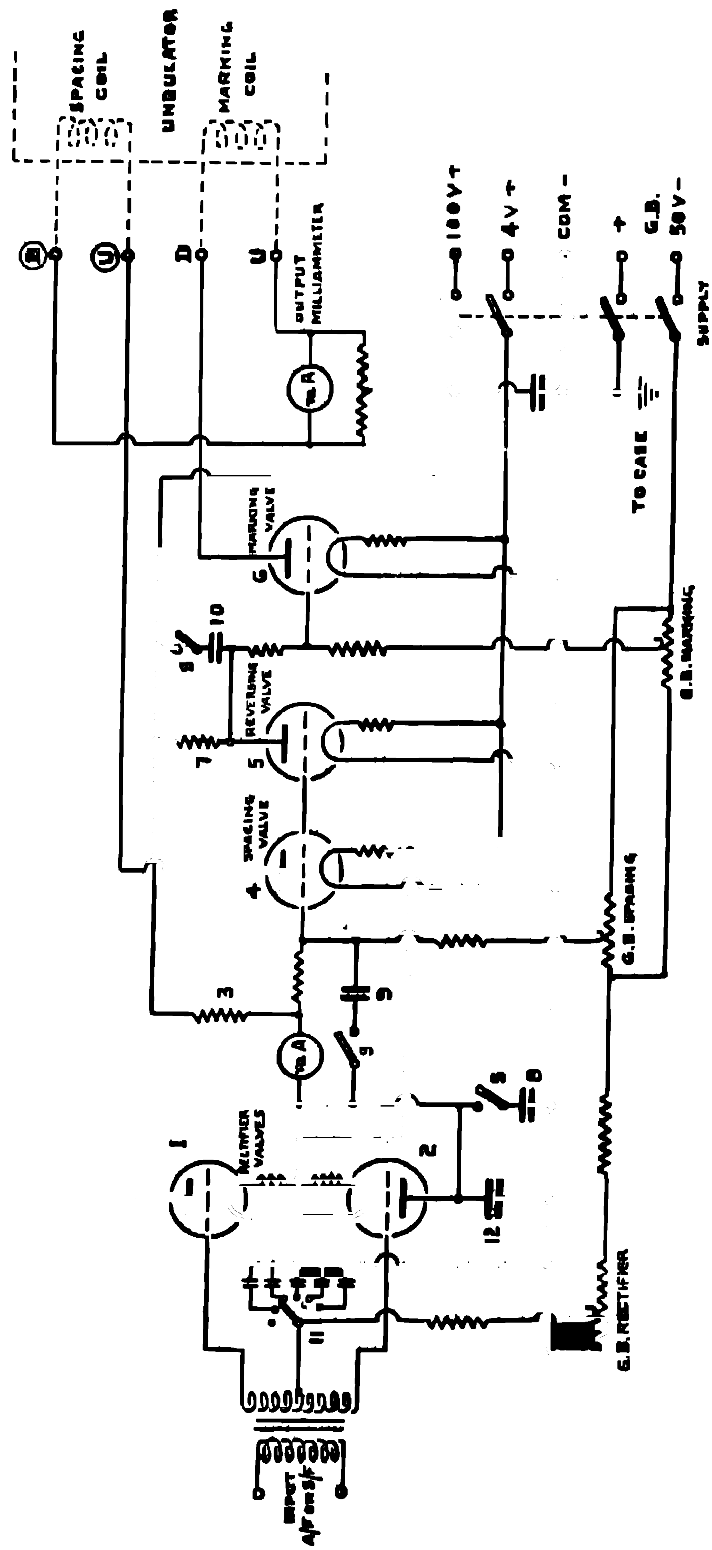

SPACING COIL
UNDULATOR
MARKING COIL
OUTPUT MILLIAMMETER
180V+
4V+
COM-
G.B.
50V-
SUPPLY SWITCH
TO CASE
MARKING VALVE
REVERSING VALVE
SPACING VALVE
G.B. MARKING
G.B. SPACING
RECTIFIER VALVES
G.B. RECTIFIER
INPUT A/F OR S/F

FIG. 51.

GENERAL CONCEPTION OF THE INSTRUMENT.—Valves (1) and (2) are operated as a *Class "B" push-pull detector combination, therefore passing no current during the no signal or spacing periods*; during the *marking periods*, current is passed which produces a corresponding IR drop in load resistance (3), which forms part of a potentiometer across the H.T. supply controlling the bias for valve (4). The latter, with valves (5) and (6), are operated as distortionless Class "A" amplifiers.

During the space period, the bias on the grid of **the spacing valve** permits current to pass sufficient to energise the spacing coil, and move the rocker arm and stylus to the space position.

During the mark period, when a signal is being received, valves (1) and (2) pass current, and the D.C. rectified component flows through resistance (3) and increases the IR drop in the latter; the grid of valve (4) becomes more negative and the current in the anode circuit of the spacing valve is reduced, and the valve may be shut down completely. Simultaneously, the same increase in negative bias is applied to valve (5), **the reversing valve.** The anode current in that valve is also reduced, and the decrease in IR drop in resistance (7) **makes the grid of valve (6) more positive by the same amount as the negative bias increase to valve (4).** This is followed by an increased current in the anode circuit of **the marking valve,** and the latter energises the marking coil and moves the writing stylus to the mark position. Since valves (4) and (6) are similar, the decrease in current energising the spacing coil is accompanied by a numerically equal increase in current energising the marking coil, the undulator constituting a mechanism operated in *pull-pull* by the D.C. amplifier. For balanced operation the instrument requires careful adjustment, an operation the technicalities of which it is not proposed fully to treat here.

It may be noted that with automatic reception it no longer becomes necessary to retain I.C.W. wave forms having modulating frequencies in the A/F range; supersonic modulation may be employed.

THE INSTRUMENT IN GREATER DETAIL.—The centre point of the input transformer secondary is connected through a variable capacity (11) to the filaments of the rectifier valves; the capacity is provided by a bank of five condensers, having values from 10 to 180 jars, operated by a switch marked **limiter time constant.** A second connection from the centre of the input transformer secondary passes through a 4-megohm grid leak to a variable potentiometer marked **grid bias rectifier,** the latter providing an adjustable negative bias to the grids of the rectifying valves.

The anodes of these valves are connected in parallel, and fed through a load resistance (3) and a milliammeter. The anode circuit is joined to the grid of the spacing valve through a potentiometer consisting of two fixed resistances, and a variable one marked **grid bias spacing,** the latter providing a small adjustment of the grid bias of the spacing valve.

The H.T. supply to the spacing valve is through one branch of a resistance connected across a differential milliammeter, marked "**output-spacing-marking,**" to one terminal of the recording undulator and hence to the anode terminal.

The grid of the spacing valve is connected in parallel with the grid of the reversing valve, the object of which is to provide values in grid potential of the marking valve which are exactly opposite to those on its own grid and on that of the spacing valve

The anode of the reversing valve is fed through a load resistance (7). The anode is also connected to the grid of the marking valve through a potentiometer consisting of two fixed resistances, and a variable one similar to that provided for the spacing valve, and marked **grid bias marking.** The anode circuit of the marking valve is also an exact counterpart of that of the spacing circuit, but uses the other branch of the differential milliammeter and the marking coil of the undulator.

Other features of interest include condensers (8), (9) and (10), these being connected together and operated by a switch marked **shaping control** (S). The effect of these condensers is either to slow up or accelerate the rise and fall of spacing and marking currents, so that the intervals of time during which the undulator holds over to mark or space are altered by a small constant amount. At slow speeds the alteration in the lengths of marks and spaces is too small to make any appreciable difference in the "shape," but at speeds over 200 words per minute, the change becomes noticeable.

With weak input signals, the rectifier valves work on the **anode bends** of their characteristics. If the working point is at (say) 8 volts negative, signals of that peak amplitude will just be sufficient fully to load the instrument. Any increase in signal amplitude above that value causes grid current to flow in the two valves alternately during successive half cycles, charging the condenser (11) to a negative potential nearly equal to the peak amplitude of the signal. This prevents any further increase in the anode current and overloading of the amplifier, and constitutes, in fact, a very effective limiter. When the signal ceases, the condenser is left charged at the potential of the signal amplitude, and the amplifier remains insensitive to any other signal of smaller amplitude until the charge leaks away through the grid leak. The time of discharge of this condenser is proportional to the CR value, and the best value of condenser is found by trial while watching the undulator tape. A large condenser is required for a slow speed signal, and a small condenser is preferable at high speeds and is essential when rapid fading is experienced in H/F reception. The limiter thus allows signals to be recorded, without distorting the shape of the morse symbol, even when the signals are much stronger than those required.

The spacing, reversing, and marking valves are worked on the part of the static characteristic between zero grid volts and the cut-off point.

Condenser (12) is a by-pass for the R/F component of the anode current of the detector valves.

METHOD OF ADJUSTMENT.—The D.C. amplifier is adjusted with no signal input. The rectifier grid bias potentiometer should first be adjusted until the anode current shown by the rectifier meter is just zero. The reversing and marking valves should then be temporarily removed, and the spacing grid bias potentiometer adjusted until the output meter indicates a suitable current on the spacing side. The rectifier grid bias potentiometer is then altered until the reading on the output meter just falls to zero ; this adjustment will alter the reading on the rectifier meter. Following this, the spacing valve should be removed and the reversing and marking valves replaced. The marking grid bias potentiometer should then be adjusted until the output meter indicates a similar suitable value on the marking side. Finally, the spacing valve is replaced and the rectifier grid bias potentiometer moved back until the rectifier meter reads zero again. During this operation the pointer of the output meter will move from a given value on the marking side, through zero, to the same value on the spacing side. The instrument will then be in the requisite state of adjustment prior to the receipt of a signal.

63. **Amplifier Noises.**—If the process of amplification is carried too far, or if an amplifier is badly designed, it will be found that many interfering noises will be heard in the telephones.

These noises may be classified as follows :—

Symptom.	*Cause.*
(1) VALVE NOISES.	
(*a*) Frying noises	Soft valve.
(*b*) Intermittent noises like atmospherics.	Bad filament.
(*c*) Clicks	Loose contacts.
(2) BAD DESIGN.	
(*d*) External noises easily picked up	Unscreened transformers and leads.
(3) HOWLING, ETC.	
(*e*) Howling in case of a note magnifier	Continuous oscillation of an audio-frequency circuit due to cross-coupling.
(*f*) Howling in case of a radio-frequency amplifier.	Generation of two radio-frequency oscillations giving an audible beat.

Symptom.	*Cause.*
(3) HOWLING, ETC.—*continued.*	
(*g*) A steady succession of clicks ..	The result of using a grid leak of too high resistance. A self-oscillation piles up a negative charge on an insulated grid faster than the leak can drain it away; the oscillation then stops momentarily and starts again as soon as the negative charge has drained off (Para 58).
(*h*) Motor boating in A/F amplifiers. The "phut-phut" noise resembling the exhaust from a motor boat.	A low frequency oscillation produced by interaction in the A/F stages.
(4) BATTERY TROUBLE.	
Crackling noises	Loose contact, or defective cells in H.T. battery.
(5) INDUCTION FROM MAINS.	
Continuous hum	Imperfect smoothing in the power rectifier unit, or faulty adjustment of the hum dinger or hum bucker control.

EXAMINATION QUESTIONS ON AMPLIFICATION.

1. Describe, in general terms, how voltage amplification is achieved in a receiver. Sketch an amplifier circuit employing "tuned anode-capacity" coupling, and explain carefully the function of each component.

   Write down an expression for the V.A.F. of the circuit, using the "equivalent diagram."

   (W/T.1 (Q), 1937.)

2. (*a*) Show why it is necessary with a three-electrode valve to have an impedance in the anode circuit comparable with the A.C. resistance of the valve, in order to obtain satisfactory voltage amplification.

   (*b*) The anode impedance of a tuned anode amplifier consists of an inductance of 400 mics. having a resistance of 10 ohms, and a capacity of 500 $\mu\mu F$. If the anode circuit is tuned to the incoming signal, what is the output voltage when a signal of 0·1 volt is applied between grid and filament of the valve, which has an amplification factor of 12 and an A.C. resistance of 30,000 ohms?

   (Qualifying Lieut. (S), 1936. Answer: 0·85 volts.)

3. Sketch the following systems of intervalve coupling used in amplifiers :—

   (*a*) Resistance-capacity; (*b*) tuned anode; (*c*) transformer.

   State their respective advantages and disadvantages, and describe clearly the action of one of them.

   (W/T.2 (Q), 1936.)

4. Compare and contrast the three main types of intervalve amplifier coupling from the point of view of V.A.F. Sketch response curves showing their relative merits over the A/F range.

   From the curves account for the popularity of the transformer coupling in the output stage of receivers.

   (Wt. Tels. (Q), 1935.)

5. (*a*) Show, from first principles, that maximum power is obtained from a triode valve when the external impedance is equal to the A.C. resistance of the valve.

   (*b*) The mutual characteristics of a triode power valve are given by the following table :—

Grid voltage		0	−2	−4	−6	−8	−10
Anode current (*m*A)	V_a = 100 volts	14·5	10	6	3	1·2	0·4
	V_a = 124 volts	20·5	16	12	7·5	4	2

   What should be the step-down ratio of the output transformer if this valve is to deliver maximum power to a loudspeaker of resistance 10 ohms?

   (Qualifying Lieut. (S), 1936. Answer: 20 : 1.)

SECTION "F."

6. A note magnifier consisting of two triode valves is to be constructed. Explain fully, with the aid of diagrams, three ways in which the two valves may be coupled together.

What are the advantages and disadvantages of each method, and what factors influence the choice of suitable components ?

(Qualifying Lieut. (S), 1936.)

7. What is the essential requirement in the output stage of a receiver ? Sketch and explain the action of a "push-pull" amplifier stage employing "mid-point biasing."

Compare the merits and disadvantages of such an arrangement with that of a stage comprising two similar valves in parallel.

(W/T.1 (Q), 1937.)

8. Make sketches showing the application of push-pull circuits to (*a*) D/F receivers, and (*b*) the power stage of receivers, stating the biasing arrangements used in each.

Explain the difference between Q.P.P. and Class "B" amplification.

(W/T.1 (Q), 1934.)

9. What are the advantages of using two valves in "push-pull" instead of a single valve for the output stage of a receiver ? Valves in "push-pull" may employ either "mid-point biasing" or "curvature biasing." Explain these terms, showing curves, and state the comparative advantages and disadvantages of the two systems.

(W/T.1 (Q), 1937.)

10. Sketch typical families of I_a/V_g and I_a/V_a characteristics for a triode used as an amplifier, and insert a typical load line for resistance-capacity coupling to the next stage.

Explain, with reference to these, how voltage amplification is obtained. What are the limitations imposed if distortion is to be avoided ?

(W/T.1 (Q), 1937.)

11. Sketch and describe the action of a transformer coupled amplifier. In the case of an ideal transformer-coupled A/F amplifier draw the equivalent circuit and derive an expression for its V.A.F.

(W/T.1 (Q), 1936.)

12. What are the forms of distortion liable to occur in amplifiers ? Compare the response obtained at audio-frequencies with the various types of coupling used in practice, and contrast their suitability for reception of W/T and R/T.

(W/T.1 (Q), 1935.)

13. An amplifier consists of a valve having an internal anode circuit impedance of 30,000 ohms and an amplification factor of 25. If an external impedance is placed in the anode circuit, what is the amplification obtainable with this arrangement if this impedance consists of—

(*a*) A resistance of 30,000 ohms.

(*b*) A coil of 5 millihenries and 25 ohms resistance, shunted by a condenser of 0·001 microfarad, the amplifier in this case being operated at the resonant frequency of this circuit ?

(C. & G. Final, 1929. Answer : 12·5 ; 21·74.)

SECTION "F."

14. In an A/F amplifier the valve has an amplification factor of 40 and an internal impedance of 40,000 ohms. It is coupled to a succeeding stage by a transformer of 1 to 3·5 ratio, having a primary inductance of 40 henries and a primary resistance of 500 ohms. The secondary load is of infinite impedance. What will be the amplification for frequencies of 40, 100, 1,000 and 10,000 cycles per second, if the primary of the transformer resonates at 10,000 cycles ?

(C. & G. Final, 1932. Answer : 34·14 ; 74·47 ; 138·2 ; 140.)

15. Sketch a suitable circuit to explain what is meant by "regenerative amplification." How does this principle assist in increasing the selectivity of a receiver ?

(W/T.2 (Q), October, 1936.)

16. Explain carefully the heterodyne method for the reception of C.W. signals, giving a sketch of a suitable circuit employing a separate oscillator. What are the advantages and disadvantages of this circuit compared with one not employing a separate oscillator ?

(W/T.1 (Q), 1937.)

17. Explain the reception of C.W. using—

(*a*) A separate heterodyne, and (*b*) an autodyne.

Give a circuit diagram in each case.

The valve used in a heterodyne circuit has $m = 40$ and $r_a = 20{,}000$ ohms. The fixed inductance in the grid circuit is 100 mics. and has a high frequency resistance of 1 ohm. If the inductance of the reaction coil is 10 mics., what is the minimum value of the coupling factor if oscillations are to be maintained at 100 kc/s. ?

(Qualifying Lieut. (S), 1936. Answer : 40 per cent.)

18. What are the causes of instability in R/F amplifiers ? What steps are taken in practice to remedy this trouble ?

(W/T.2 (Q), 1935.)

19. Enumerate the types of receivers suitable for the reception of H/F signals.

With the aid of a circuit diagram, and necessary curves, describe the action of a separate quench receiver for I.C.W. signals. What considerations govern the choice of quenching frequency ?

(W/T.1 (Q), 1937.)

20. Explain, with the aid of a schematic diagram, the superheterodyne principle of reception.

What points in the design of such a receiver are of great importance. Why is a second oscillator valve unnecessary when receiving a modulated wave ?

(W/T.2 (Q), 1935.)

21. In an audio-frequency amplifier, the valve has an internal impedance of 30,000 ohms and an amplification factor of 40. It is coupled to the succeeding stage by a transformer having a ratio of 1 : 3 and a primary inductance of 30 henries. The secondary load is a resistance of 50,000 ohms.

Neglecting the internal capacitances of the valve and the self capacitance, resistance, and leakage of the transformer, what will be the amplification for frequencies of 100 and 10,000 cycles per second ?

(C. & G. Final, 1937. Answer : 54·2 and 77·9.) (Cf. Paragraph 19.)

SECTION "F."

22. Define (*a*) the decibel, (*b*) the neper.

A low frequency amplifier has a gain of 56 decibels. The input circuit is of 600 ohms resistive impedance and the output is arranged for a load of 10 ohms. What will be the current in the load when an alternating potential of 1 volt is applied at the input?

A transmission line has an equivalent of 7 decibels; what is the equivalent of this in nepers?

(C. & G. Final, 1937. Answer: 8·1 amps., 0·806 neper.)

POWER SUPPLIES.

1. **General Requirements.**—Considered as a whole, any W/T installation requires power for its receivers and transmitters. The method of supply can be separately considered under these two headings, although the problems are not separate ones and it may be more economical to consider both of them jointly.

In the case of receivers, those in the Service, in common with many others, are usually designed to employ a high tension D.C. supply of either 50 or 100 volts; power valves in loudspeaker systems frequently need D.C. voltages of the order of 400 volts. The low tension supply is usually at 4 volts. This demand for power is one which could be met by means of batteries. For an installation having several receivers, this is a method which has many disadvantages. Batteries take up space, are very heavy, and need constant care and attention to prevent deterioration and to maintain them in a state of efficiency. Moreover, it is a very inconvenient method of producing D.C. potentials of the order of 400 volts. Their chief advantage lies in their excellent voltage regulation characteristic. Due to the low internal resistance of an accumulator, the maximum variation of voltage with varying load is usually about 10 per cent. It is not easy to design other forms of power supply for receivers with as good voltage regulation as this. In spite of this difficulty, various very satisfactory power units can be produced, and it is fast becoming common practice to use rectified and smoothed A.C. supplies for the production of all of the D.C. voltages that may be required.

In the case of transmitters, low power sets using L/F may require a high tension voltage at the transmitting valve anodes of the order of 100 or 200 volts. This could be supplied by accumulators or small D.C. generators. If, however, ranges of hundreds or thousands of miles are required, it is necessary to increase the H.T. voltage to thousands of volts. In the section on "Valve transmitters" (Section "K"), it is shown that the power delivered to the oscillatory circuit is proportional to the product of the emission current of the filament and the voltage applied to the anode; with correct circuit adjustment it therefore follows that, for a given value of current, the greater the voltage on the anode, the greater will be the power delivered to the oscillatory circuit. The limiting factor to an indefinite increase in either or both of these quantities is provided by the anode rating of the valves employed, as with (say) a 66 per cent. efficiency circuit, half as much power as is given to the oscillating circuit is dissipated in the form of heat at the anodes.

Steady potentials of the order of 10,000 volts may be produced by :—

(*a*) A motor generator giving a direct current output at very high voltage;

(*b*) A motor alternator and transformer, the output from which is rectified, normally by the use of thermionic valves.

The motor generator method has advantages as regards efficiency and space occupied, but under ship conditions it is generally not so reliable as the low voltage motor alternator and transformer. In addition, the commutator sparking on a high voltage motor generator may give rise to serious interference with reception, unless a static suppressor is carefully fitted at the source. The majority of Service transmitters use a rectified A.C. supply.

There are many different rectifying devices. Rectifying valves are discussed in the section on "Thermionic valves" (Section "B"). For power rectification, employing valves, it is usual to use diodes, although it is obvious that triodes could also be employed. The general principles of a power rectifier will be made clear by considering the cases of the half-wave and full-wave valve rectifiers.

2. **The Diode Half-Wave Rectifying Circuit.**—Fig. 1 illustrates the simplest type of valve rectifying circuit. It consists of an alternator and step-up transformer, in the primary circuit of which there is a switch which may be called the "signalling key." The rectifying valve is generally known as a "U" valve. The heating current of the filament is supplied by means of a motor-alternator and step-down transformer. Condenser C is of large capacity, and is a storage or reservoir condenser; it is also called a **smoothing condenser,** since it is the source of the rectified D.C. supply.

The circuit also shows supply leads to the anode and filament of the oscillating valve, or load circuit.

When the switch in the step-up transformer primary circuit is made, an alternating voltage of large amplitude is applied to the anode of the rectifying valve. If the filament is alight, a current only flows through the valve when there is a positive P.D. from anode to filament, *i.e.*, during the positive half-cycle of alternator voltage. During the negative half-cycle, the anode is negative to the filament and no current can flow. When current is flowing, electrons are being transferred from the filament to the anode, *i.e.*, from one plate of condenser C to the other, and so a P.D. is established across the condenser. Assuming that the lower plate in Fig. 1 is earthed, the potential of the upper plate and the filament thus rises during the positive half-cycle of alternator voltage, and so diminishes the part of this half-cycle during which the anode is positive to the filament, *i.e.*, the time during which the condenser is being charged. If the transmitting circuit (or load circuit) is in operation, current is always flowing out of the smoothing condenser to supply the steady anode current taken by the transmitting valves, the condenser being, in fact, equivalent to the H.T. battery in a receiving circuit. At the beginning of the rectifying process, the charge received by the condenser, while current flows in the rectifying valve, is greater than the charge it loses to the transmitting valves during one cycle of alternator voltage. As the P.D. across its plates rises, however, and the charge received by it decreases, a state of equilibrium is eventually reached when the anode of the rectifying valve only remains positive to the filament for such time

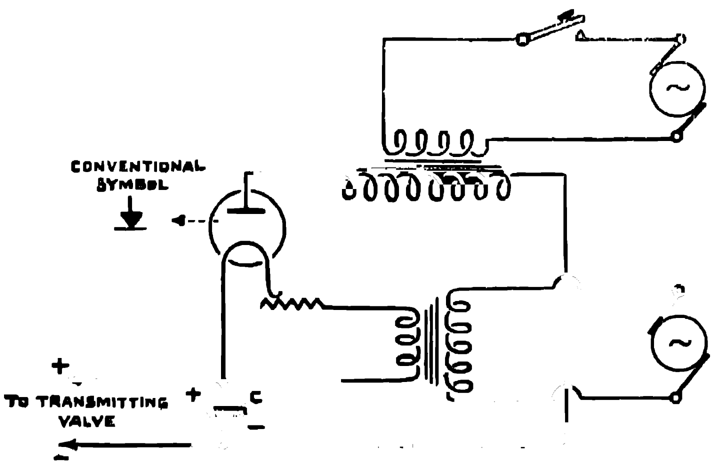


FIG. 1

as will suffice to make up for the loss of charge to the transmitting valves per alternator cycle. The P.D. across the condenser then maintains a sensibly constant value, apart from a slight ripple due to the fact that it is charged discontinuously, while its discharge through the transmitting valves is continuous.

If the mean condenser P.D. were to rise above this equilibrium value, the charge it would receive would not be sufficient to make up for the losses to the transmitting valves, and the P.D. would fall again. If it fell below the equilibrium value, the current, flowing through the rectifying valve for a longer time, would charge the condenser faster than it discharged, until the equilibrium state was again reached.

If the transmitting circuit is not in operation (*i.e.*, if there is no load), the P.D. across the plates of the condenser rises until it is equal numerically to the peak value of the A.C. supply. When this state has been reached, assuming that there are no losses in the condenser due to leakage, the anode of the valve will not be positive with respect to the filament at any time during the cycle of alternator voltage. An electrostatic voltmeter connected across the condenser would then indicate

the peak value of the A.C. supply ; "peak voltmeters" working in this way are familiar instruments in any laboratory.

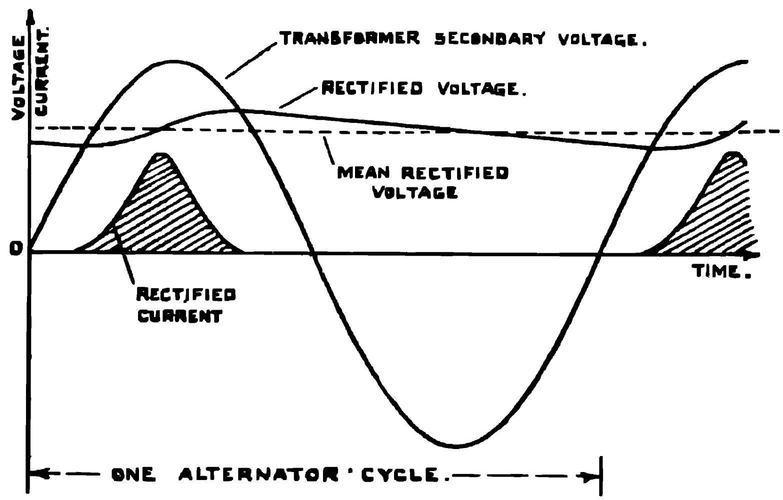
The electron path should be noticed. Starting from the positive condenser plate, electrons travel from filament to anode of the rectifying valve, and through the transformer secondary to the filaments of the transmitting valves, thence through these valves back to the positive condenser plate.

In early transmitting circuits, use was made of rectifying valves of comparatively small filament emission, with the result that the rectifier current reached saturation almost as soon as the anode-filament P.D. became positive, and remained at its saturation value practically until the anode potential again fell below that of the filament. This sytem was obviously inefficient, for once the anode-filament P.D. was sufficient to give saturation, no increase in rectifier current was obtained during the time that the anode-filament P.D. was greater than this value. With the demand for ever increasing power in the aerial, it has become necessary to work the alternator and transformer as nearly as possible at their maximum output. Hence, the filaments of rectifying valves are now designed to give sufficient emission to prevent saturation ever being reached during the rectifying cycle, and the current through the rectifying valve increases as long as the anode-filament P.D. increases.

With the circuit shown in Fig. 1, rectified current only flows during part of the positive half-cycle. The process is therefore called **half-wave rectification.** The action is illustrated in Fig. 2.

The transformer secondary voltage is shown for simplicity as a simple sine wave. This is not the case in practice, owing to the irregular power output during a cycle. This point is considered below (para. 9).

The current flowing through the rectifying valve when the anode is positive to the filament obeys the 3/2 power law (Section "B") as indicated in the figure. The shaded area represents the charge acquired by the smoothing condenser during one cycle of alternator voltage. The mean



HALF – WAVE RECTIFICATION.

FIG. 2.

condenser P.D. is shown by the dotted line, while the continuous line indicates how its P.D. actually varies during a cycle. The discharge to the transmitting valve goes on steadily, and when the condenser is not being charged, its P.D. falls uniformly according to the steady anode current in the transmitters. During charging, the condenser P.D. rises approximately as indicated, the rate of charge being a maximum when the anode-filament P.D. of the rectifying valve is a maximum, if the lag due to the inductance of the transformer secondary is neglected. This H.T. voltage ripple is one of the factors leading to frequency variation in C.W. transmitters (Section "K"). It can

easily be seen that its amount varies inversely with the capacity of the smoothing condenser, since the smaller this capacity the greater is the change in P.D. produced by a given charge $\left(V = \frac{Q}{C}\right)$. The greater the capacity of the smoothing condenser, the longer does it take to discharge after the circuits are switched off. If primary signalling is being used, the effect would be to maintain the transmitter oscillating after the key in the primary of the H T transformer is broken, thus effectually setting a limit to the speed of signalling (K.55) The size of the condenser is usually fixed by such considerations as its cost, and the amount of ripple that can be allowed.

3. **Full-Wave Rectification.**—This is the system normally employed in practice. The circuit is shown in Fig. 3 Two rectifying valves are used, often included in the same envelope—

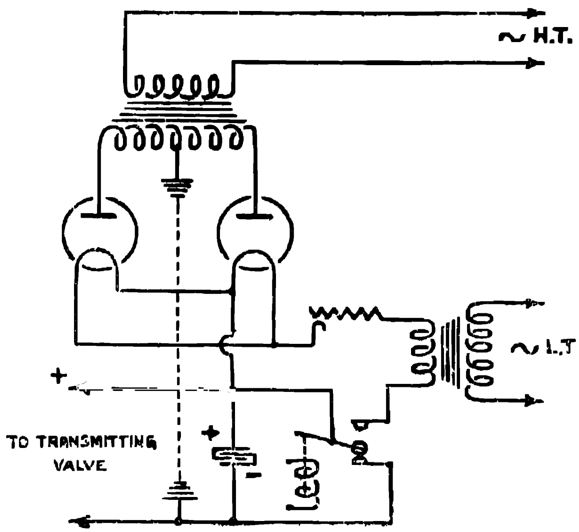
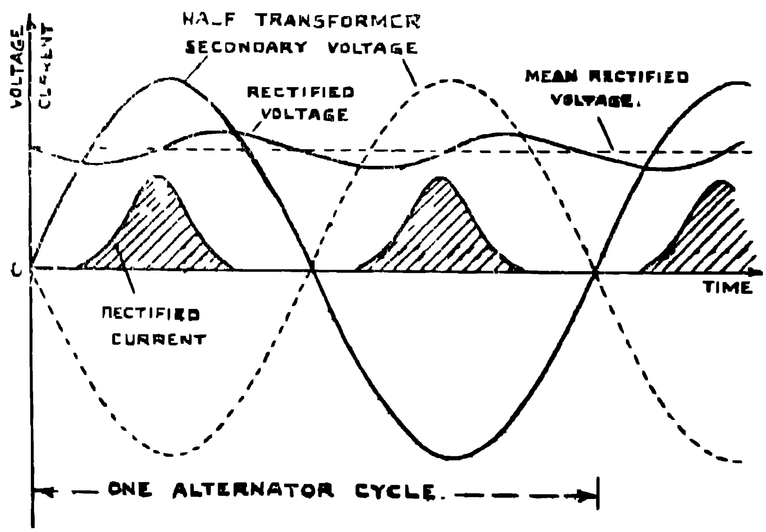


FIG. 3.



FULL - WAVE RECTIFICATION.

FIG. 4.

Fig. 15. Their anodes are connected to the terminals of the transformer secondary, the centre point of the secondary being earthed and therefore common with the negative plate of the smoothing condenser. Thus the alternating potential applied to the anode of each valve is half the total secondary voltage, and maximum positive potential on one anode corresponds in time to maximum negative potential on the other. Consequently, current flows in the two valves during alternate half-cycles, the rectifying action of either valve being the same as that described above. The process is illustrated in Fig. 4.

It should be noted that, strictly speaking, the name applied to this method of rectification should be "bi-phase half-wave rectification," since each half of the transformer supplies current for half a cycle. By common custom, the method of rectification described above is called "full wave rectification," although that title may be more appropriately reserved for cases such as Fig. 10 (*d*), (*f*) and (*g*).

The smoothing condenser receives a charge during every half-cycle of alternator voltage, instead of once per cycle as in half-wave rectification. The condenser voltage ripple is therefore less pronounced and occurs at twice the frequency of the alternator.

4. **Efficiency of a Rectifier Unit.**—An exact calculation of the efficiency is complex, as will be realised from the descriptive account of rectification given above. It must take into account the $\frac{3}{2}$ power law of variation of rectified current with anode-filament P.D., the distortion of the transformer secondary voltage wave-form, the smoothing condenser rectified voltage ripple, and the impedance of the transmitting circuit. Even an approximate calculation presents difficulties which render it outside the scope of this book, but some approximate figures will be quoted below.

The efficiency of the rectifying system is the ratio of its power *output* to the transmitting circuit to its power *input*, which consists of the main alternator power output and the power expended in the filament in producing the emission. The principal power loss arises in the heat produced at the anodes of the rectifying valves.

The efficiency increases with the ratio of H.T. rectified voltage to transformer secondary voltage amplitude. It is therefore an advantage to rectify at as high a value of this ratio as possible. For this reason, a rectifying circuit is arranged to give as low a peak anode-filament P.D. as possible; this is the reason why some types of rectifying valves have long narrow anodes. In one Service case the peak anode-filament P.D. was 3,000 volts during the positive half cycles. The peak anode-filament P.D. during the negative half cycles was, of course, much greater than this; this is usually called the "**inverse voltage,**" and in the above practical case, with a rectified voltage of 10,000 volts and a secondary amplitude of 13,000 volts, the inverse voltage amounts to 23,000 volts, and the valve insulation must be capable of standing up to this requirement. The total secondary P.D. in the above case, however, would be 26,000 volts. This gives a ratio of H.T. rectified voltage to secondary P.D. across each valve of $\frac{10}{13}$ or $0 \cdot 77$.

Decreasing the peak P.D. would, necessarily, increase the efficiency of rectification. The transformer secondary voltage, however, is more or less fixed for a given alternator and transformer, and any reduction of the alternator voltage usually spoils the alternator "regulation" (the rise and fall of voltage with load), consequent on the production of low magnetic densities in the iron. Any reduction in voltage may best be effected by means of an auto-transformer arrangement of the primary winding.

In the above practical case there was a smoothing condenser ripple of about 4 per cent. using a condenser of capacity $1 \cdot 5\ \mu F$.

In paragraph 2 it was observed that the percentage high tension ripple varied inversely with the capacity of the smoothing condenser. In more precise terms it may be said to vary with the ratio X/R, where X is the reactance of the smoothing condenser, and R is the effective resistance presented by the load. This ratio will be small when X is small, or when R is big; when R is infinitely big, the rectifier is working under no load conditions. For reasons already partly referred to in paragraph 2, the value of the capacity is usually chosen to compromise between conflicting requirements. The use of too large a smoothing capacity involves pulses of charging current with

peak values which are too high for the anode rating of the valve. This sets an upper limit to the size of the condenser, and we must compromise between the requisite degree of smoothing and the size of the rectifying valve.

5. Elementary Discussion of the Design of the Smoothing Unit—Percentage Ripple. —By making certain simplifying assumptions, it is possible to arrive at a design criterion for a smoothing unit consisting of a condenser, as in Figs. 1 and 3. With reference to Fig. 2, the variation in rectified voltage across the condenser is a wave form which may be simplified into a series of saw teeth, as shown in Fig. 5.

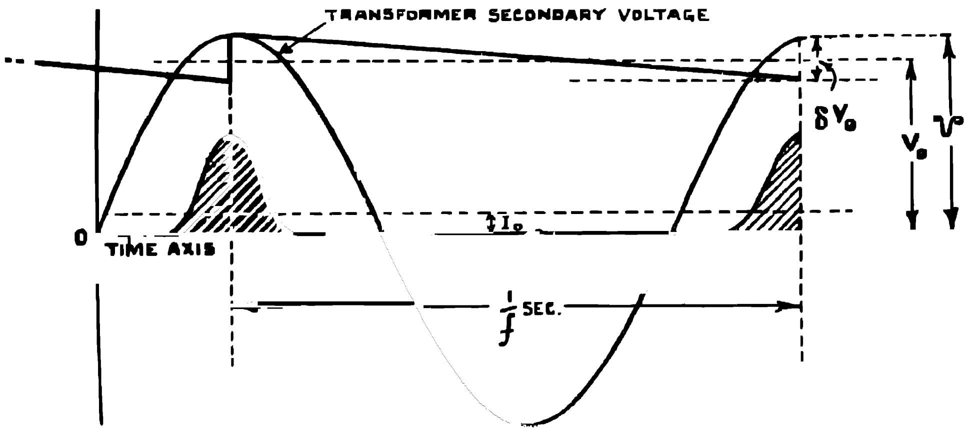


FIG. 5.

When the load is relatively small, it is possible to make the following good first approximations :—

(*a*) The mean rectified voltage (V_0) may be put equal to the peak value of the transformer secondary voltage ($\mathcal{V}$).

(*b*) The maximum variation in voltage across the condenser (δV_0) is due to the steady load current (I_0) discharging for a period of time (δT). This period of time may be taken to be roughly $1/f$, where "f" is the frequency of the A.C. supply to the transformer primary. The value of the load current is roughly given by

$$I_0 = \frac{V_0}{R}, \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

where R represents the effective value of the load on the rectifier. From simple condenser theory we now have—

Quantity of electricity = current × time = Capacity × voltage change. In symbols, with the notation of the "Calculus"—

$$\delta Q = I_0 \delta T = C \delta V_0$$

and from (1)

$$\frac{V_0}{R}\frac{1}{f} = C\delta V_0.$$

$$\therefore \qquad \text{Per cent. ripple} = \frac{100\,\delta V_0}{V_0} = \frac{100}{RfC}. \quad \ldots\ldots\ldots\ldots\ldots \quad (2)$$

From (2) it follows that the percentage ripple will be small when the denominator is big. In order that a C.W. transmitter may preserve a constant frequency, it is essential that the H.T. supply should be as free from ripple as possible; an alternator producing A.C. at 500 cycles/second is, therefore, more suitable for use in a rectifier unit designed for use with a transmitter than one producing (say) 50 cycles A.C. Partly for this reason, many Service transmitters use alternators working at 500 cycles/second. Using single phase A.C. at 50 cycles, it would be exceedingly difficult to provide the requisite degree of smoothing needed by a C.W. transmitter, though it would be entirely feasible to do it using a 3-phase supply at that frequency.

Considering a numerical example :—

If $R = 10^4$ ohms
$f = 500$ cycles/sec.
$C = 1\ \mu F$.

$$\text{Per cent. ripple} = \frac{10^6 \times 10^2}{10^4 \times 500 \times 1} = \frac{10^4}{5} = 20\ \%.$$

In the case of a full wave rectifier, the percentage ripple would be 10·0 per cent. (A 4μF condenser would give better results.) In practice, these percentages would be smaller.

Equation (2) may be modified by using $X = \frac{1}{\omega C}$ and $\omega = 2\pi f$

$$\frac{\delta V_0}{V_0} = \frac{\omega}{R f \omega C} = \frac{2\pi X}{R}. \qquad \text{(3)}$$

Moreover, the average value of the rectified voltage (V_0) is given by—

$$V_0 = \mathcal{V} - \frac{\delta V_0}{2} = \mathcal{V} - \frac{2\pi X V_0}{2R} \doteqdot \mathcal{V} - \frac{\pi X \mathcal{V}}{R}$$

$$= \mathcal{V}\left(1 - \frac{\pi X}{R}\right) \text{ and when } \frac{X}{R} \text{ is small}$$

$$V_0 \doteqdot \frac{\mathcal{V}}{1 + \frac{\pi X}{R}}. \qquad \text{(4)}$$

From (3) it follows that the percentage ripple will be small when the ratio X/R is also small. This form of the relation, together with (4), also stresses the point that any given percentage change in R without a corresponding change in X, will produce a considerable alteration in the percentage ripple, and in the value of the mean rectified voltage V_0. In order to minimise large percentage changes in R, it is advisable that there should always be some small load on the rectifier; it is partly for this reason that sometimes a high resistance (*e.g.*, 150,000 ohms) is connected permanently across the smoothing condenser. The percentage change from 150,000 ohms to 15,000 on load is a smaller percentage change than from (say) 1 megohm to the same value on load. It should be noted that the connection of a high resistance across the condenser is also necessary, in the Service, in order to conform with certain safety regulations; it produces a discharge path to earth when the rectifier is not in use.

From (4), when X/R is small in comparison with unity, we have

$$V_0 \doteqdot \mathcal{V}.$$

6. Overheating of the Anodes.—Overheating of the RECTIFIER anodes indicates that the rectifier filament emission is too low, and it should be increased. This increases the charge received by the smoothing condenser under the given conditions, and so raises the equilibrium P.D.

across the said condenser. The efficiency of rectification is therefore increased and the power dissipation at the rectifier anodes is diminished, with a corresponding diminution in anode temperature.

Overheating of the TRANSMITTING valve electrodes, if the anode tap is correctly adjusted, indicates that more power is being developed in the transmitting circuit than the anode rating of the valves will allow. The output from the rectifier unit should then be lowered by reducing the main alternator voltage. This reduces the H.T. rectified voltage proportionally and so diminishes the power supply to the transmitting circuit.

7. **Filament Arrangements.**—As the rectifier filaments are at a high potential to earth, they and their heating circuit must be carefully insulated. This is most easily accomplished by using an alternating heating current from a motor-alternator and step-down transformer, as shown in Figs. 1 and 3. It is a comparatively simple matter to provide efficient insulation for the transformer secondary winding.

In full wave rectification, it may be convenient to provide a separate rheostat or choke for each filament to control its emission in case the characteristics of the two rectifying valves differ.

In Fig. 3 there is also shown a mechanically operated switch, called the rectifier switch. In the position shown, this short circuits the smoothing condenser when the rectifier is not in use. This is most important, as the charge remaining on the condenser may easily give a fatal shock to the operator. In the other position, the rectifier switch makes the filament heating circuit.

In many more modern rectifying units, the former function of the switch is performed by a discharging resistance permanently connected across the condenser; the value of the resistance is high and so constitutes only a small permanent load, other advantages of which have already been pointed out in paragraph 5.

8. **Equalising Arrangements.**—In cases where large powers are being dealt with, the arrangement shown in Fig. 3 of tapping off the H.T. rectified supply from *ONE SIDE* of the two rectifying valve filaments, is unsuitable.

The filament is generally arranged as a loop, and in the case of Fig. 3, one side of the loop would be carrying a much greater proportion of the rectified current than the other. The extra current would raise the temperature and increase the emission from one half of the loop and this half would wear out sooner than the other, giving the filament an unduly short life.

It is preferable to arrange the rectifier filament circuits as in Fig. 6, where there is an EQUALISER COIL across each filament, and the high tension lead to the smoothing condenser is connected to the mid point of each coil.

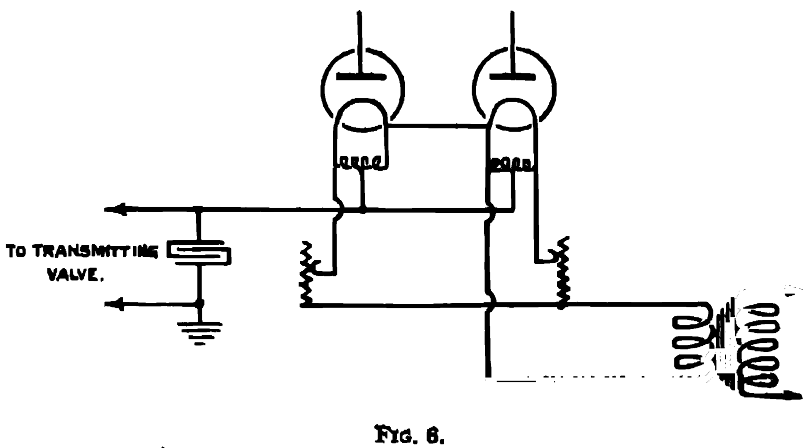


FIG. 6.

The equaliser coil has a high inductance, and consequently takes very little of the filament heating current. The flow of H.T. rectified current, however, is in opposite directions in the two halves of the coil, which is consequently non-inductive to it.

In this way, paths of equal conductivity are provided for the rectified current in both sides of the filament simultaneously, equalising the current distribution in the two halves of the filament loop.

An alternative method of achieving the same result would be to tap the mid point of the secondary winding of the filament transformer, connecting the filament to the two ends of the winding, and the smoothing condenser between the mid point tapping and earth. The above method of using chokes, has, however, the advantage of eliminating the unbalanced condition which would result, with the centre transformer tapping system, if the filament regulating resistance is included in one limb only.

EQUALISING RESISTANCES may be used instead of equalising coils; Fig. 7 is a circuit showing the full wave rectifier employing equalising resistances (1) with R/F by-pass condensers (2) connected across them. The resistances should be large enough to constitute a negligible load in comparison with the filament load; in one Service case the filament is supplied at 22 volts, and across it are connected in series two equalising resistances each of 25 ohms.

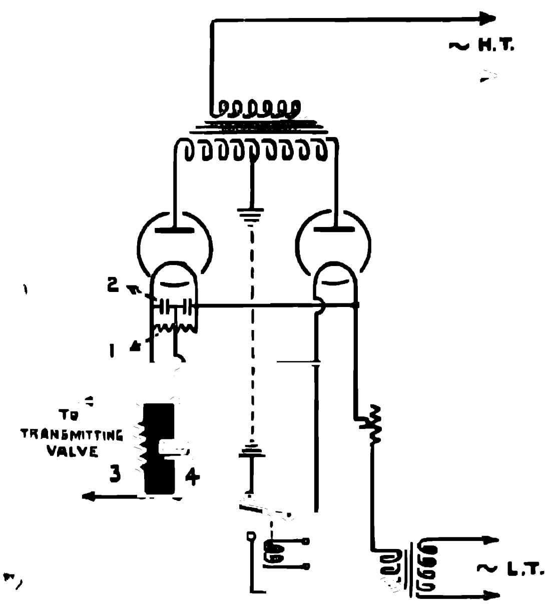


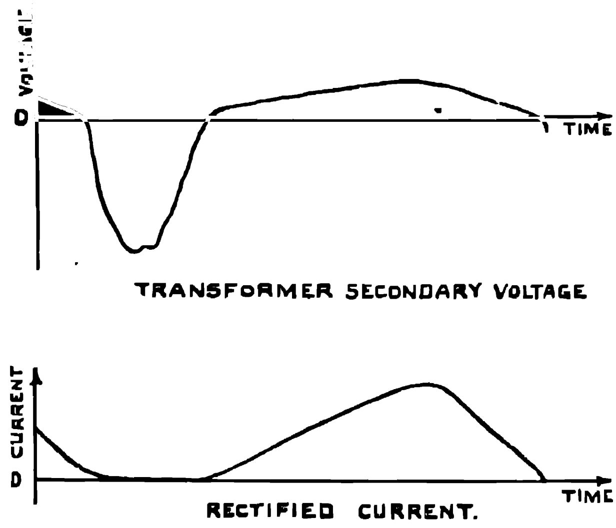
FIG. 7.

Fig 7 also shows the discharging resistance (3) permanently connected across the smoothing condenser.

Finally, it may be observed that the necessity for equalising arrangements is not restricted only to rectifying valves but applies equally to all valves when large powers are being dealt with.

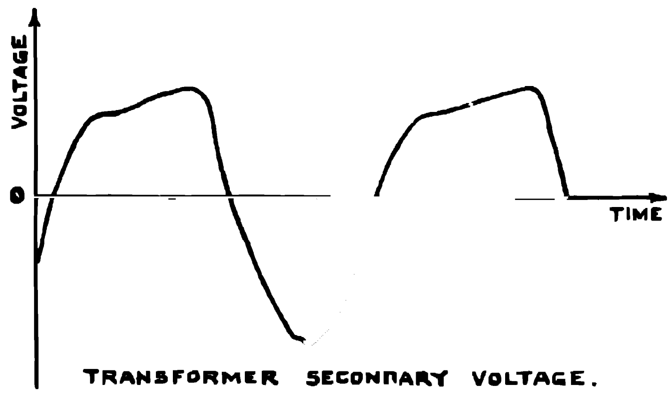
9. **Transformer Secondary Voltage.**—In the above discussion it has been assumed that the transformer secondary voltage follows a simple sine curve, but the variable load presented by the rectifying valves renders it impossible for this to be the case in practice.

In the analysis of half-wave rectification given above, the current flows through the rectifying valves in pulses corresponding to parts of the positive half cycles of secondary voltage, and no current flows during the remainder of each cycle. The effect of this variable load is to distort the secondary wave, *i.e.*, to introduce harmonics of the alternator frequency. The rectified current flowing through the secondary winding also has a mean value in one direction, since all the pulses of current are in the same direction through the winding. This produces a steady magnetisation of the core, and therefore distorts the secondary voltage wave, as explained in Volume I. The effect, in practice, is to increase the time during which the current flows in the rectifying valve, but to reduce the maximum value of rectified current. Experimental curves of the anode-filament P.D. and the rectified current during one cycle, are shown in Fig. 8. It will be observed that there is a considerable second harmonic in the voltage waveform. This is largely due to the transformer iron distortion, and is not nearly so pronounced in full wave rectification, as shown by the experimental curve of Fig. 9. The rectified current pulses from the two valves then flow in opposite directions through the secondary winding, and so the two half-cycles of the voltage waveform are more similar in shape, *i.e.*, even harmonics are reduced, and may be eliminated if the two valves are identical and have the same filament emission.



HALF-WAVE RECTIFICATION.

FIG. 8



FULL-WAVE RECTIFICATION

FIG. 9.

10. **Metal Rectifiers.**—There are many other "one way devices" with properties similar to those of diode valves, enabling them to be used as rectifiers.

"Metal rectifiers" are very frequently used instead of valve rectifiers to produce the H.T. supply for receiving and low power transmitting circuits when the primary electrical supply is alternating. The metal rectifier operates on the same principle as the crystal detector. The most usual form is a copper disc, one face of which is oxidised by a special heat treatment. The contact between the film of copper oxide and the metallic copper behaves like a crystal and cat's-whisker, or crystal couple. A voltage applied in the direction from oxide to metal gives a much greater

current than the same voltage applied in the opposite direction—from metal to oxide. Alternatively, the metal rectifier may be compared with a diode in which the oxide is the anode and the metal is the cathode or filament.

The flow of current during rectification develops heat in the disc, and, as the temperature of the disc increases, the difference between its conductivities in opposite directions is diminished, *i.e.*, its efficiency is decreased ; further, overheating may cause chemical changes which permanently impair the action. There is hence an upper limit to the voltage which may be applied across any one disc, and, for higher voltages, two or more rectifiers in series must be used. To increase the flow of rectified current for a given applied voltage, the appropriate number of rectifiers may be connected in parallel.

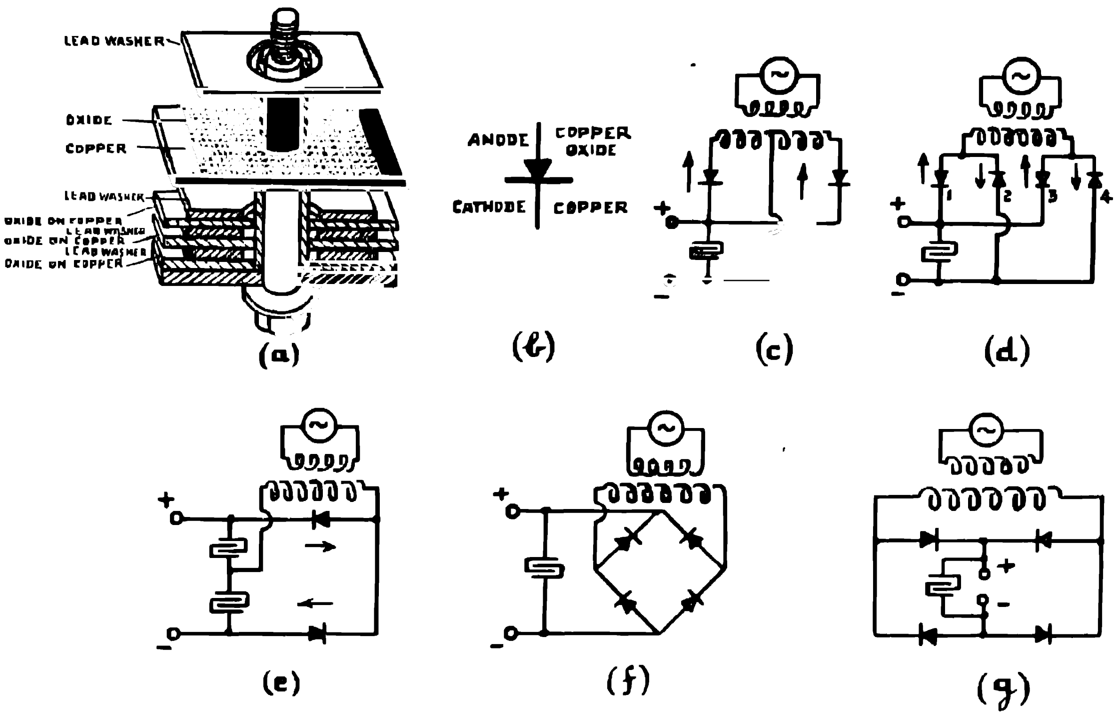


FIG. 10.

The ordinary rectifier unit consists of a number of rectifier discs, Fig. 10 (*a*), mounted in series on a screwed bolt and insulated from it by a sleeve. The discs are separated by lead washers, and conducting spacers and large metal fins are interposed alternately between discs to dissipate the heat developed.

Provided a metal rectifier unit is not overloaded, its life should be very much longer than that of a valve rectifier.

Typical " full-wave " metal rectifier circuits are shown in Fig. 10, although Fig. 10 (*c*) is not strictly " full wave " (H.3). The arrowhead in the " bench mark " symbol for a metal rectifier indicates that direction of an applied voltage for which it is conducting, and hence also indicates the direction of conventional current flow through it. The electronic flow is in the opposite direction.

Fig. 10 (*c*) corresponds exactly to Fig. 3, the valve full-wave rectifier. In Fig. 10 (*d*) an arrangement is shown whereby the centre tapping on the transformer secondary winding may be

dispensed with. Rectifiers (1) and (4) are conducting during one half-cycle of applied alternating voltage, and rectifiers (2) and (3) during the next half-cycle. For the same transformer secondary voltage, the output voltage is twice that given for the circuit of Fig. 10 (*c*). Figs. 10 (*f*) and (*g*) show two alternative ways of drawing the circuit of Fig. 10 (*d*); the symmetrical arrangement of Fig. 10 (*f*) is very commonly seen.

Fig. 10 (*e*) shows a circuit giving a rectified voltage approximately twice the value of the transformer secondary voltage. The two rectifiers, connected in opposite directions to the same secondary terminal, behave as half-wave rectifiers during alternate half-cycles, and provide rectified voltages in the same direction across their respective smoothing condensers. These condensers are in series and so their voltages are additive. Voltage doublers are frequently used in X-ray and other low power circuits where high voltages are required and the inverse voltage would be too high for the ordinary circuits; with suitable arrangements, voltages may be produced with as large a value as 400,000 volts.

The film of copper oxide on the copper is so thin that the arrangement acts as a capacity with a resistance in parallel. The high capacity of this rectifying device has presented a difficulty hindering its use in R/F rectifying circuit.. The difficulty has, however, been overcome in various ways, one of which is by the use of rectifying surfaces of small area. The metal rectifier is an almost perfect linear detector, and can be made to take the place of a valve rectifier in most receiving circuits, provided that it is preceded by stages of R/F amplification. The copper oxide rectifier gives good results with input voltages above a minimum of 3 volts, and so functions well as the (say) second detector in a superheterodyne circuit.

11. **The Mercury Vapour Rectifier.**—Diodes are now almost solely used for rectifying alternating current. Small double diodes in glass bulbs are used for supplying currents of the order of 0·1 amp. for high tension purposes in receiving apparatus, while larger ones with silica or water-cooled envelopes are used for delivering several milliamps at many thousands of volts to transmitting sets, in the way that has been described in the foregoing paragraphs. In the case of all of the ordinary hard diodes, there is a considerable loss of power inside the valve owing to its high internal resistance. In the case of the example quoted in Section "B," for a current of 2 amperes, the anode voltage is of the order of 1,000 volts, which represents an A.C. resistance of approximately 500 ohms. The "regulation characteristic" of an ordinary diode is very poor.

In recent years a very successful type of soft diode has come into use for power rectification.

The "gas" remaining in the valve is mercury vapour in equilibrium with the liquid. When the cathode is heated it emits electrons which are attracted to the anode in the ordinary way; as they pass across the tube frequent collisions with mercury molecules will occur, and if the velocity of impact is great enough the molecules may be split into ions. In practice it is found that for a given soft diode of this kind, the ionisation potential is roughly constant at about 15 volts. With that P.D. between anode and cathode, the electrons can just acquire the requisite speed between collisions with mercury molecules in order completely to ionise the vapour. When the P.D. is below 15 volts the current is very small and the resistance of the tube is quite high. At the ionising potential, the current suddenly increases vastly and the resistance drops almost to zero; at the same instant, a blue glow permeates the tube.

For comparative purposes Fig. 11 shows the characteristic curves of a soft diode and a hard diode respectively. For an anode current of 200 milliamps, there is a drop in P.D. across the hard diode of 105 volts, the drop across the mercury vapour diode being only 15 volts. It is obvious that a valve with a characteristic of this kind represents an approach to the ideal rectifier; the low value of the internal resistance implies that the "voltage regulation" will be very good.

When ionisation occurs, a part of the increase in anode current is due to the assistance of the *ionic carriers* of electric charges, the process being analogous to electrolytic conduction. It has, however, been shown that the relatively slowly moving ions do not constitute a large part of the anode current, the principal cause of this increase being the neutralisation of the negative space charge near the cathode by means of positive ions. When this has been done, electrons will be

drawn to the anode as fast as they are emitted from the cathode, provided that the P.D. is equal to about 15 volts. Roughly, it may be said that the function of the "gas" is the neutralisation of the space charge.

The presence of relatively heavy positive ions is attended with certain disadvantages. Positive ions falling on the cathode surface are liable to cause its disintegration. It is found, however, that provided the ions do not fall through a potential difference of more than 22 volts they have not a seriously injurious effect on the surfaces. It is, in fact, possible to use ordinary oxide coated emitting surfaces, and with anode voltages of between 15 and 20 volts positive with respect to the cathode, it is easy to have an anode current constituting the entire filament emission.

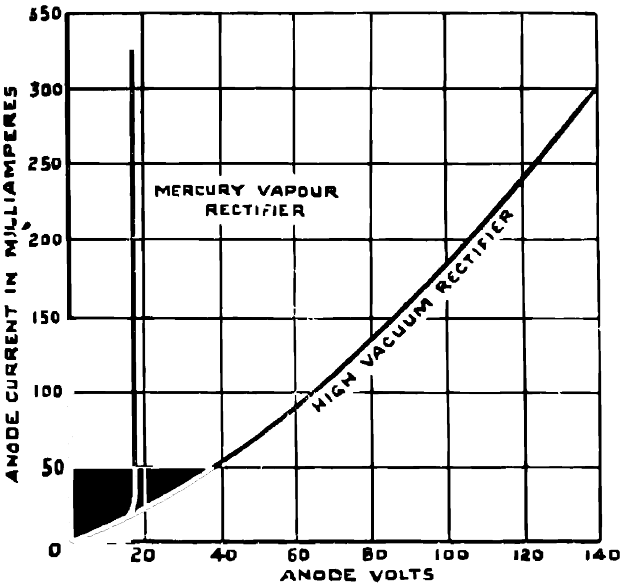


FIG. 11.

Soft valves suffer from the drawback—for Naval purposes—of delayed action on starting. The cathodes require 20 seconds or so to heat up to the point of maximum emission, and it is dangerous to apply H.T. for at least half a minute, since there may not be sufficient electron emission to produce ionisation of the gas and to prevent the voltage across the tube from rising above the value which will disintegrate the filament or cathode surface. Even before general ionisation takes place, there are some positive ions left within the tube, and to prevent injury to the cathode it is essential that these should not fall through a potential difference greater than 22 volts. Premature application of H.T. must, therefore, be avoided; this delay sometimes constitutes a disadvantage. In commercial practice, special thermal delay action switches are in common use, in order to avoid applying the H.T. too early.

The circuits already described for half and full-wave rectification, and voltage doubling, may be applied equally well to the mercury vapour rectifier. It is important that the load resistance in the circuit should always be sufficiently high to prevent the appearance of a P.D. of more than about 15 volts across the valve. An accidental short circuit across the load would apply the full P.D. of the alternator secondary across the tube with fatal results to the latter; an ordinary diode rectifier might stand this treatment, since the first action would only be the heating of the anode. Care must also be taken that the maximum "inverse voltage" does not rise as high as the "sparking voltage" through the low pressure mercury vapour tube; this is somewhat lower than would be the case if mercury vapour were not present.

12. **Voltage Regulation in Rectified A.C. Supply Units.**—When any source of power—such as an accumulator or a dynamo—is put on load, the terminal P.D. falls, the relation between the terminal P.D. and the load being called the "voltage regulation characteristic." The numerical value of the fall will depend upon the relative value of the load resistance and the internal resistance of the source of power. The fall in P.D. is sometimes referred to as the "lost volts" in the power unit.

In paragraph 5, equation (4) shows very approximately the way in which the value of the rectified voltage (V_0) varies with the load R in the case of a rectifier with a smoothing condenser. In the case of most rectifiers and smoothing units, approximate calculations can be made of all aspects of their performance with the exception of the voltage regulation; this is a matter which

presents considerable difficulty. The excellent voltage regulation of accumulators and mercury vapour rectifiers has been referred to in paragraphs 1 and 11. If it were desired to operate a variety of receivers from a rectified A.C. supply, using a single SUPPLY UNIT, the voltage regulation characteristics of the supply unit would become of great importance. In the case of both H.T. and L.T. supply units, the internal resistance of the rectifiers commonly used is very high compared with that of a battery of accumulators, and the consequent bad voltage regulation with load is made worse by the resistance of the smoothing chokes, which are usually included as well as a smoothing condenser.

For comparison, Fig. 12 represents the voltage regulation characteristics of a mercury vapour rectifier, and a hard diode rectifier, each working in similar full-wave rectifying circuits. The relative steepness of the two sets of curves demonstrates the superior regulation characteristics of the mercury vapour rectifier.

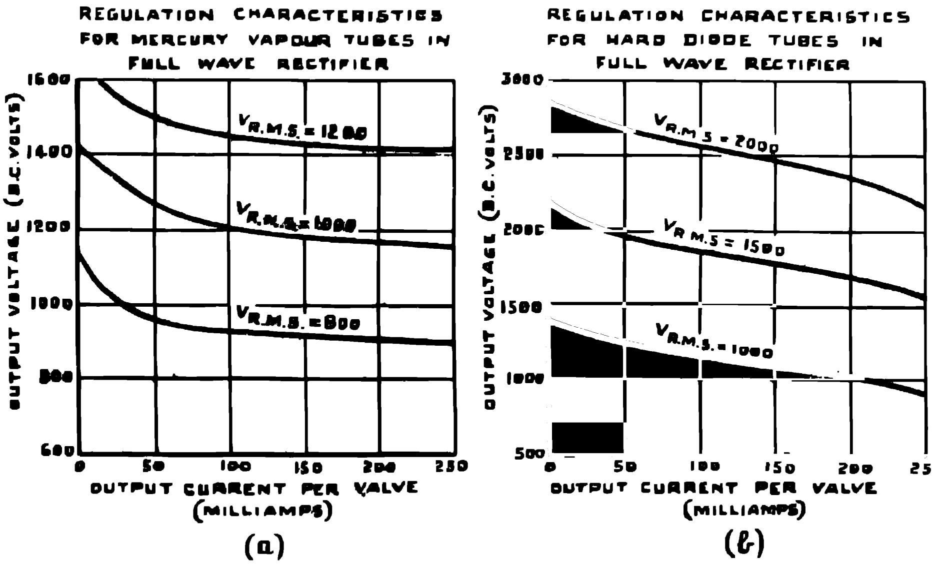


FIG. 12.

There are three ways in which the voltage output from a rectifier unit may be maintained within prescribed limits :—

(*a*) A rectifier may be used of much greater maximum output than the biggest load to be taken from it, the rectifier being loaded with a permanent load so that the variations in load may be small.

(*b*) A constant load may be maintained on the rectifier equal to the maximum load ; this may be done automatically or by switching.

(*c*) The input A.C. voltage to the rectifier may be increased as the output load increases.

In one Service supply unit, giving both H.T. and L.T. direct current outputs, method (*b*) is used to stabilise the H.T. supply, the L.T. supply being stabilised by method (*c*). In most cases it would be economically unsound to employ method (*a*).

13. **L.T. Voltage Stabilisation.**—Fig. 13 (*a*) represents the circuit details of a method of low tension voltage stabilisation, in which the A.C. input voltage to the rectifier is automatically increased as the external D.C. load current increases, so as to maintain the load voltage within the prescribed

limits. In this method, considerable over-compensation of the rectifier output D.C. voltage is necessary because of the resistance of the filter circuit. For example, if the filter circuit has a resistance of 1 ohm, the output D.C. voltage from the rectifier must increase from 4 volts to 6 volts as the load current increases from 0 to 2 amps, in order to maintain the load voltage at 4 volts. To reduce the total over-compensation required, it is necessary to make the D.C. resistance of the output filter circuit as low as possible.

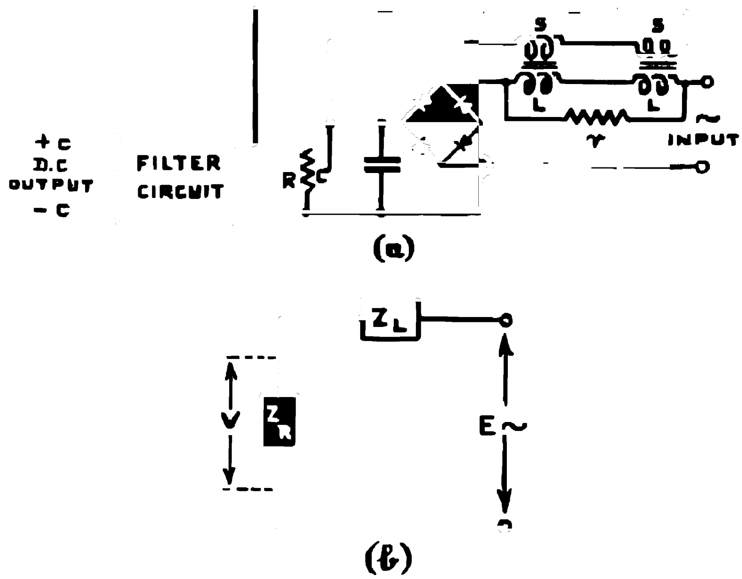


FIG. 13.

The A.C. input to the rectifier passes through two identical chokes (L) on closed iron cores. The D.C. load current passes through the polarising windings (S) on the chokes. As the load current increases, the iron cores of the chokes approach saturation with consequent reduction of the permeability, and hence the inductance of the chokes (Volume I); the impedance of the chokes therefore decreases as the load increases.

The input resistance of the rectifier also decreases as the load current increases, but if the percentage decrease in impedance of the chokes is greater than that of the rectifier, the A.C. voltage input to the latter will increase.

The bridge rectifier circuit is made up of metal rectifiers, the impedance of which varies considerably between no load and full load In one case, it was found to be 130 ohms at no load, compared with 20 ohms for a load of 0·1 amp. at 4 volts, and 3·9 ohms for a load of 2·0 amps. at 4 volts. To avoid the necessity of compensating for the large variations of rectifier impedance with very light loads, a semi-adjustable load resistance R is connected across the output terminals of the rectifier in order to take a small permanent load (paragraph 5).

The polarising windings (S) in which the D C. rectified output flows, are arranged so that the induced 50 cycle voltages cancel each other (a "hum bucking" arrangement). Although this prevents any E M.F. at the supply frequency (50 cycles) being introduced into the output circuit from the polarising windings, even harmonics of the supply frequency do not cancel, and as a result of the distortion of the supply waveform by the saturated cores of the chokes, large voltages at these frequencies are induced into the output circuit. The attenuation of the L.T. filter circuit must be adequate to prevent interference with reception from this source. The diagram of a complete filter circuit in a rectified A.C. supply unit is shown in Fig. 15.

The circuit of Fig. 13 was designed to give at the output terminals a ripple-free D.C. supply

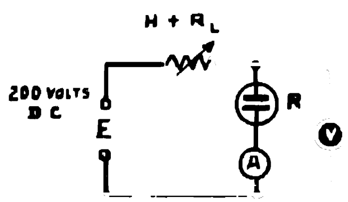
voltage which remained within the limits 3·6 to 4·2 volts for any load current up to 2·0 amperes. The stabilising resistance *r* (about 20 ohms) prevents "motor boating"; since the P.D. across L decreases as the load increases, "negative resistance" may exist.

This circuit could be analysed mathematically from the equivalent diagram of Fig. 13 (*b*) in which E = the applied A.C. voltage, V = the voltage input to the rectifier, Z_L = the impedance of the chokes at the supply frequency, Z_R = the rectifier impedance at the supply frequency. From Ohm's law we have

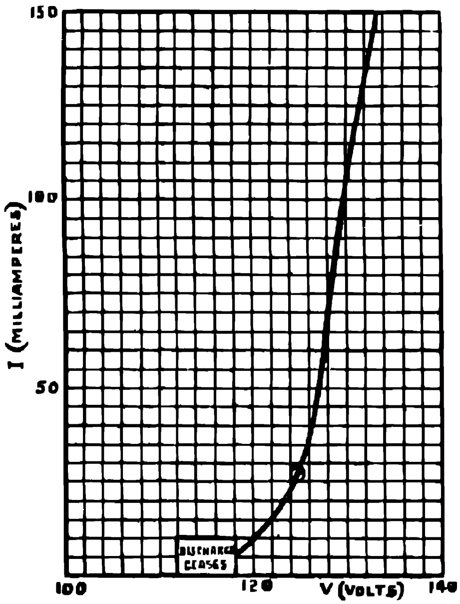
$$V = E\frac{Z_R}{Z_L + Z_R},$$ and V is the voltage which it is desired to maintain constant.

Z_L and Z_R are each separately dependent on the D.C. load current. From this point, the remainder of the analysis is left as an exercise for the mathematician.

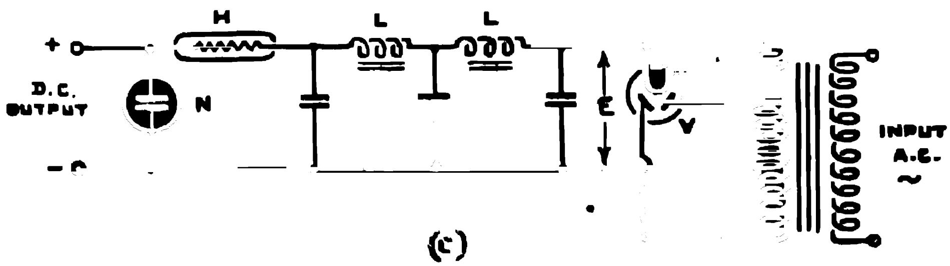
14. **H.T. Voltage Stabilisation.**—Fig. 14 (*c*) gives the circuit details of an automatic method of H.T voltage stabilisation using a compensating load in parallel with the external load; the former INCREASES and decreases as the external load DECREASES or increases, in such a measure as to maintain a constant load on the rectifier and hence a constant output voltage. In



(a)



(b)



(c)

FIG. 14.

the case described in this paragraph, the glow discharge tube constitutes the compensating parallel load. In paragraph 15, the H.T. voltage stabilised circuit therein described uses a triode valve to maintain a constant load on the rectifier.

Glow discharge tubes are the now familiar neon tubes, containing the gas, mixed with certain others, at a pressure of about 10 mm. of mercury. To understand their action, consider the experiment represented in Fig. 14 (*a*). If V represents an ordinary voltmeter, no appreciable current will pass through the tube until H is reduced, so that the potentiometer formed by H and V allows the voltage across V to become equal to the " striking potential " of the tube—about 140 volts. At that point the gas suddenly ionises, a change which is accompanied by a rapid fall in effective resistance of the tube. After ionisation, the voltage across the tube is never small ; the " running voltage " is lower than the striking voltage, and is only a few volts higher than that at which the discharge ceases and recombination of the ions takes place. The voltage at which the discharge ceases is called the " extinction potential," and the P.D. across the tube is always between that and the running voltage.

It may be noted that if this experiment is conducted in a darkened room, it is usually possible to detect a glow in the neighbourhood of the anode when the P.D. across the tube is well below that necessary to cause ionisation ; actually, a minute current flows through the tube. It may also be noted that if the voltmeter V were an electrostatic voltmeter, the tube will always strike provided H is finite.

In many respects, the glow discharge tube resembles the mercury vapour rectifier discussed in paragraph 11. This is evident from an inspection of the characteristic curve of Fig. 14 (*b*), from which it is seen that, on ionisation, the effective resistance of the tube decreases very rapidly from about 400 ohms to about 70 ohms in the running condition. The " effective resistance " of the tube varies with the P.D. across it, and may be determined from the slope of the characteristic curve at any desired point. In the normal running condition, the tube may be considered to be working towards the bottom of the straight portion of its characteristic (as indicated), in which condition the resistance is about 70 ohms.

The graph shown in Fig. 14 (*b*) represents the characteristic curve of a typical commercial tube rated to stabilise a load of 0-60 milliamps at 125 volts. Its stabilising action may be appreciated by assuming an equation for the straight line portion of the characteristic

$$I = \frac{V}{r} - b,$$

where $\frac{1}{r}$ is the slope of the curve, and is therefore large, and where b is a constant.

This may also be written in the form $V = rI + rb = rI + E$.

From this it appears that the output voltage available across the tube is the sum of a large CONSTANT term " E " (a back E.M.F.), and a small VARIABLE term " rI." In this variable term, " r " is the reciprocal of the slope of the characteristic and, therefore, represents the effective resistance of the tube, which normally has the low value of about 70 ohms. The current I is kept small by having the resistance H large. The large value of the latter resistance is thus an essential feature of the stabilising device, quite apart from the fact that it must always be large in order to limit the maximum current through the discharge tube to its rated maximum value. Too large a value of the discharge current will produce an arc and destroy the tube.

The method of connecting the discharge tube voltage stabiliser is shown in Fig. 14 (*c*). As the external load increases the output volts tend to decrease, causing a large decrease in current through the tube. If the external demand ceases, the rectifier output volts tend to go up, and the current through the tube increases very greatly for a small increase in voltage. The nett effect of connecting the tube in parallel with the external load is to maintain a nearly constant current demand on the rectifier. In one practical case, a variation in load current of 50 milliamps produced a variation in output voltage of approximately 2 volts ; this may be compared with a variation of 50 volts in the case of a full wave hard diode rectifier.

In order to ensure that the discharge tube will strike when the supply outfit is switched on,

with the maximum load already connected to its terminals, it is necessary that the resistance H should have a much lower value initially than finally. This condition may be achieved by using a tungsten filament lamp as the resistance H. The advantage of this may be made clear by an example. Let the neon tube have the following characteristics :—

Striking voltage	=140 volts.
Running voltage	=100 volts.
Maximum working current	= 60 milliamps.
Resistance of choke	=200 ohms.
Maximum load	= 50 milliamps, corresponding to a resistance of 2,000 ohms.

Then if H is a 220 volt, 25 watt tungsten filament lamp, with a cold resistance of 300 ohms and a hot resistance at 80 volts of 1,300 ohms, the voltage across the load when the supply is switched on is—

$$V = E \times \frac{2000}{2500}, \text{ from which } E = 5/4\,V. \qquad \text{(para. 13)}$$

Hence, if V is equal to or greater than 140 volts, E must be equal to or greater than 175 volts.

If E is made 190 volts, the tube will strike with the heaviest load connected, and the maximum current through the tube (with no load) will be 60 milliamps. If the resistance of H were 1,300 ohms initially, the tube would not strike if an output load of less than 4,200 ohms were in circuit with the supply voltage of 190 volts.

The commercial " Stabilovolt " valves represent an improvement on the simple neon tube with two electrodes. Over a considerable distance between the two electrodes within the simple tube there is a regular fall of potential, suggesting that the tube virtually has the properties of a potentiometer. It has been found that by arranging additional electrodes between the anode and cathode, advantage may be taken of this potentiometer effect. In the Marconi Co.'s tube, the anode and cathode consist of co-axial cylinders one within the other, and the other electrodes are cylindrical grids suitably spaced between the inner and outer cylinders. A tube of this nature makes it possible simultaneously to stabilise a number of different D.C. voltages ; in one practical case, the voltages available are 280 volts, 210 volts, 140 volts and 70 volts.

15. **A Comprehensive A.C. Supply Unit.**—Fig. 15 represents the circuit details of a supply unit working from a single phase A.C. input of 230 volts at 50 cycles/sec.

On the output side appears—

(*a*) 4 volts A.C., maximum current 6 amps., for the filament heating of indirectly heated valves.

(*b*) 4 volts D.C., maximum current 2 amps., for the filament heating of directly heated valves.

(*c*) A choice of 50 volts or 100 volts D.C., maximum current 0·1 amp., for the H.T. supply to receivers.

The circuit details of the L.T. voltage stabiliser have already been described above. The initial adjustment of the L.T. output voltage is carried out by means of the semi-adjustable resistance R, of 18 ohms maximum value ; this is set so that the required voltage regulation characteristics are obtained. The input voltage will be about 9 or 10 volts.

The H.T. supply is obtained from a full-wave rectifying valve, the filament and anodes of which are fed by windings from the transformer. An unsmoothed D.C. output of about 180 volts at 100 milliamps can be obtained from the rectifying valve. This is smoothed by a two-stage low-pass filter, using chokes (1) and condensers (2) in conjunction with reservoir condenser (10). The voltage is reduced by means of the lamp resistance (3), and the adjustable resistance (4) of 600 ohms maximum value. The voltage at the 100 volt D.C. output terminals is stabilised by means of either a discharge tube voltage stabiliser, or a triode valve. The alterations in circuit to enable either method of stabilisation to be used are made by means of links and sockets.

In the case of the discharge tube, the links put the tube in circuit in the way which has already been described above. In that case the lamp (3) may be a 100 volt 16 watt metal filament lamp, and the resistance (4) is adjusted initially so that the current through the discharge tube, with no external load, is about 80 milliamps.

The triode stabilising circuit employs a triode of low A.C. resistance and high mutual conductance, which fits the same holder as is employed for the discharge tube. In this position, the links connect the anode to the positive H.T. output terminal and remove the earth connection from the grid, connecting the latter to a tapping point on the potentiometer formed by resistances (5), (6) and (7), connected between the 100 volt positive output terminal and the common negative. The filament of the stabilising valve is heated from a winding on the transformer which has a centre tap connected to earth.

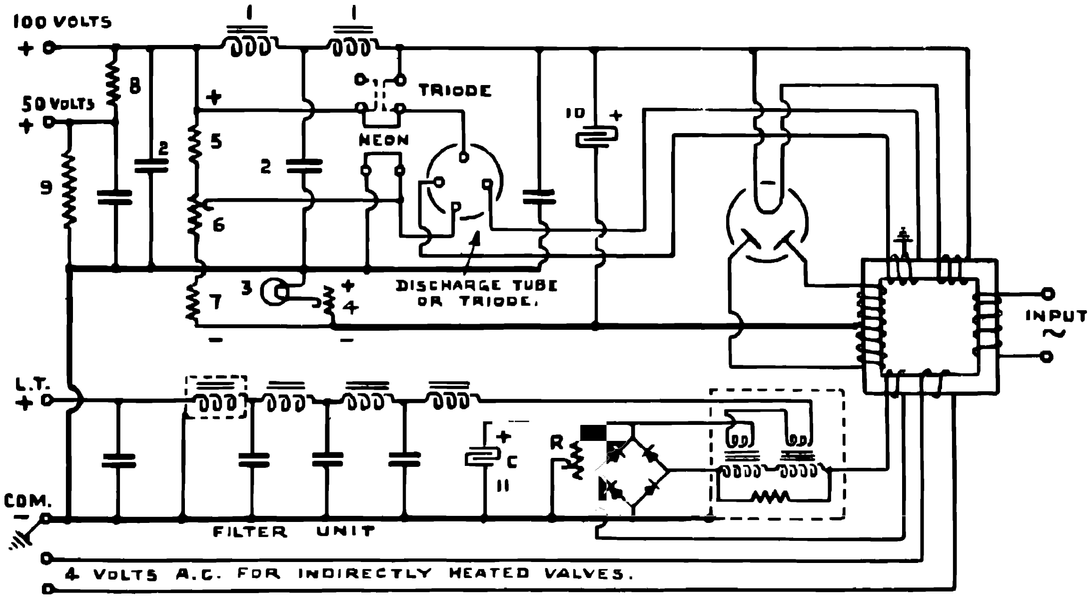


Fig 15.

The method of operation of the stabilising valve is as follows :—

The bias on the grid of the valve is the result of the NEGATIVE bias set up by the current I passing through the resistances (3) and (4), equal to the valve current plus the external load current, and the POSITIVE bias given by the potentiometer. The nett bias, with no external load, is zero or slightly negative. When an external load is connected, the current I increases ; this makes the grid of the stabilising valve more negative. The voltage drop in the chokes (1) increases this negative bias still further. Consequently, the stabilising valve takes less current, and the increase in total current I is much less than it would be in the absence of the triode. The additional or load current is supplied at the expense of the valve current, thereby tending to maintain the total current at a constant value. The fall of output voltage with the increase of load current is thereby much reduced. In one practical case, the load variation of 50 milliamps produces a voltage variation of from 20 to 25 volts. Using the triode stabiliser, resistance (3) may conveniently be a 220 volt 16 c.p. carbon filament lamp.

The initial adjustment of the H.T. output voltage is made by means of variable resistance (4) and the grid bias potentiometer (6), so that the output H.T. voltage, with no external load, is 110 volts with a current of 80-100 milliamps. passing through the stabilising valve.

The 50 volt H.T. output terminal is fed from the 100 volt terminal through resistances (8) and (9), constituting a potentiometer across the H.T. supply.

The rectifying condensers in the model have the following values: Condenser (10), 500 volts test, 4μF.; condenser (11), 1,000μF. (electrolytic type).

16. **Central W/T Power System.**—In the foregoing paragraphs, the provision of power for receivers and transmitters has been considered as if it were a separate problem. In many cases it may be so, but in the case of large W/T installations, such as those in ships, it may well be more economical, and better in other ways, to avoid a piecemeal solution of the problem. This would involve the adoption of some "central power supply system," the details of which would depend upon requirements.

A convenient form of central power might be 3-phase alternating current at 50 cycles per second.

For the use of receivers, low power transmitters, and for some auxiliary services, a single phase A.C. supply at 230 volts could be obtained from that source. Rectification could be undertaken as requisite in the vicinity of the receiver or other apparatus requiring D.C. input.

For the use of high powered transmitters, it would be necessary to use the 3 phases of the central power supply. The main reason for this is that very much less smoothing would be required than in the case of single phase. The employment of 50 cycles single phase A.C. for high power transmitters would involve the use of smoothing equipment of prohibitive size and weight. This matter has already been referred to in paragraph 5, and also provides a reason for the very common use of A.C. at 500 cycles in connection with the H.T. supply to transmitters. Moreover, many transmitters producing I.C.W. rely on the frequency of the main alternator for the production of the necessary modulation, *e.g.*, a 1,000 cycle note in the case of a full-wave rectifier working at 500 cycles. If, however, a valve modulating circuit is employed instead of the main supply, modulation may be undertaken at any desired frequency, and there is no longer any valid reason for the retention of 500 cycle A.C. generators.

In the case of a large W/T installation a central power supply system has many advantages, among which may be included:—

(*a*) Reduction in the number of machines required.
(*b*) Only one type of alternator would be needed.
(*c*) Considerable saving in space and weight would result—an important matter in ships.
(*d*) With the general use of alternating current in the neighbourhood of the W/T installation, interference to W/T reception due to commutation noises would be considerably reduced.

With any such system it would be essential to have very efficient voltage control arrangements.

POWER SUPPLIES.

EXAMINATION QUESTIONS.

1. Give a circuit diagram of a full-wave thermionic rectifier complete with smoothing circuit suitable for supplying high-tension direct current to a valve transmitter.
(C. & G. I., 1933.)

2. Describe a full-wave thermionic rectifier suitable for converting a single-phase, low-tension alternating current to high-tension direct current for the anode supply of a C.W. transmitter. Mention particularly any parts of the apparatus which must be insulated to withstand high potentials.
(C. & G. I., 1928.)

3. Describe fully a method of obtaining low-tension and high-tension current supplies for a broadcasting receiver from A.C. mains.
(I.E.E., October, 1926.)

VALVE TRANSMITTERS.

1. **Historical Note ; C.W. Transmission.**—In Section "A," paragraph 1, reference has already been made to the extreme obsolescence of the spark transmitter. The excessive interference produced by the radiation of damped wave trains, combined with a growing demand for radio-telephony, served as the stimulus to the development of C.W. transmitters. The first of these involved the use of either a high frequency alternator, or the Poulsen arc. The production of C.W. oscillations by means of the thermionic valve was, in the early stages, restricted in its use to local heterodyne oscillators and low power transmitters. The development of the valve transmitter has, in some way, been proportional to the rate at which the power-handling capacity of the valve has increased, the latter having run concurrently with receiving valve development. The power handling capacity of the early valves was very small, and valve transmitters compared unfavourably with existing methods of producing C.W.

Some idea of the progress of the development of valve transmitters—slow at first, but now proceeding fast—may be gleaned from an inspection of the following list of approximate dates :—

1913—The use of "soft" valves was generally appreciated in receivers.

1915—Successful telephony experiments between America and Paris. The need was recognised for higher powered valves, with consequently higher vacuum. 40 to 50 watts was not a big enough "anode rating" to achieve long distance telegraphy at the low frequencies then in use.

1916—Ship installations were still mostly arc transmitters. New ships began to be fitted with valve sets. The production of "hard" valves had commenced, and, in June of that year, signals were successfully sent between Portsmouth and Gibraltar. In that case the Tungsten anodes of the valves were worked at 2,000 volts, and a current of about 10 amps. was obtained in an aerial having a total resistance of 5 ohms.

1917—The Marconi-Osram Co. commenced the production of transmitting valves for the Admiralty, and the glass valve with an anode dissipation of 150 watts was made in this year. French authorities issued thousands of small valve transmitters and receivers for use on the battlefields.

1919—The obsolescence of the spark transmitter definitely began. C.W. was in general use in big stations, usually produced by the Poulsen arc or H/F alternator. Silica valves were introduced, those with an anode dissipation of 1 kW being quite common.

1920—Thermionic valves were still somewhat limited in power and only used in small transmitters, since it was uneconomical to use a large number of small valves in parallel.

1922—Experimental valves rated at 100 kW. had been produced, but the 500 watt glass bulb transmitter was still the most common. Crystal control of frequency began to develop from this year. The B.B.C. was founded and erected eight stations within the year.

1923—Demountable valves were produced by Holweck, and two rated at 10 kW. were installed at the Eiffel Tower.

1924—The beginning of the H/F era. Successful telephony between Poldhu and Sydney. The invention of the copper-glass seal enabled a great increase to take place in the anode rating of valves ; the anode became accessible and could be water jacketed in order to cool it.

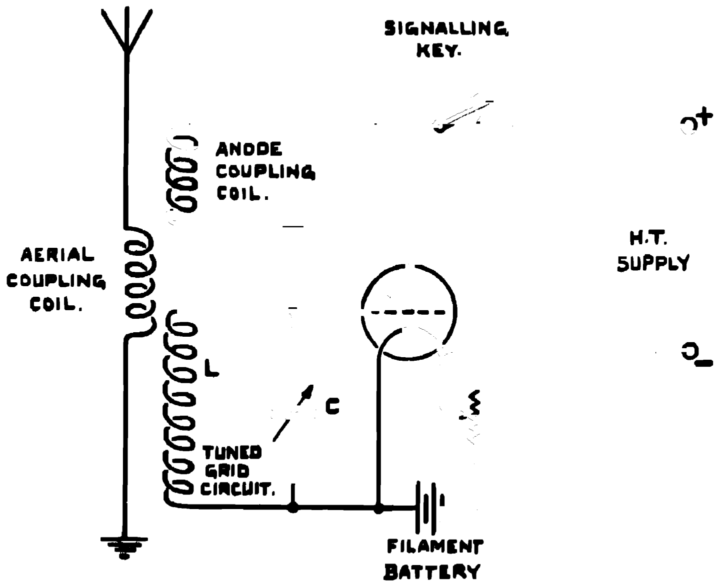
1927—The G.P.O. opened a transatlantic telephone service using the "single side band" system. The power of the transmitter was about 100 kW., and the final stage at first included thirty 10 kW. water-cooled valves.

1931—A 500 kW. demountable valve was installed experimentally in the single side band transmitter.

1932—Commencement of the B.B.C. Empire H/F beam broadcasts.

1933—The British telephone subscriber able to communicate with about 95 per cent. of the world's telephones, by means of the numerous radio-telephone links which have been established.

✓ 2. **The Simplest Valve Transmitter.**—It should now be recognised that the æther waves set up by the flow of an oscillatory current in an aerial circuit are of the same nature as the current. If the current is damped, the æther waves are similarly damped. If an undamped oscillatory current can be produced, then undamped æther waves will be sent out from the aerial.



SIMPLE VALVE TRANSMITTER.

FIG. 1.

A simple valve circuit capable of producing an undamped oscillatory current is shown in Fig. 1. The filament heating circuit and the H.T. battery between anode and filament are already familiar. It is to be noted that the positive terminal of the H.T. battery is not fastened directly to the anode, but to one end of an inductance, the other end of which is connected to the anode. This does not affect the permanent deficit of electrons on the anode under such conditions, but it does mean that the electrons flowing through the valve to the anode to supply the deficiency (when the grid allows them), then flow through the inductance in the anode lead on their way back to filament through the key and the battery. Between the grid and filament of the valve there is connected an oscillatory circuit.

Until the key is pressed no current will flow through the valve, but there will be an emission of electrons from the filament, their number depending on the heating current; the emitted electrons form a cloud in the neighbourhood of the filament called the **space charge** (B.21).

When the key is pressed, the anode is made positive to the filament and electrons start to flow across the valve. Since there is an inductance in the anode lead, the current will not rise instantaneously to the value it would have if the positive terminal of the battery were connected directly to the anode. A current whose value is changing (increasing) flows through the inductance; if there were no grid circuit attached to the valve, the current could be assumed to grow exponentially up to its final steady value.

Since, however, there is an oscillatory circuit attached to the grid, the inductance of which is magnetically coupled to that in the anode lead, any change in magnetic flux produced in the latter must be accompanied by some reaction produced in the former. Lenz's law tells us, shortly, that the direction of any reaction in the grid inductance coil must be such that it tends to counter the change of flux. Since the **anode and grid coils of Fig. 1 are each wound in the same direction,**

an electron current moving *upwards* through the anode coil produces flux in the grid coil which can only be countered by an electron current moving *downwards* through L. In the oscillatory circuit, the bottom plate of C therefore becomes negative, acquiring a surplus of electrons, while the top plate suffers from an increasing deficit of electrons, and therefore becomes positively charged. This implies that the grid becomes positive with respect to the filament, a fact which further tends to increase the flow of electrons from filament to anode in the valve, and so further to increase the downward flowing electrons through L ; a self-supporting system appears to have been set up, since the reaction of the system is to produce an effect tending to support the initial cause.

The current in the anode coil does not go on increasing indefinitely, nor does the condenser C continue to be charged in a way which makes the top plate more and more positive. The anode current tends to become steady at the value it would have if there were no grid circuit ; due to the mutual action of the grid coil L, and the inertia of the oscillatory circuit, the anode current overshoots its steady value, rising to some value higher than its value under static conditions. Ultimately, the anode current starts to fall, and an increasing electron current commences to flow in the *downward* direction through the anode coil. For the same reason as before, this must produce an electron current which rises *upwards* through L, thus making the grid potential pass from positive values, through zero, to a negative potential with respect to the filament Since a positive potential on the grid assisted an electron current moving upwards through the anode coil, the negative potential on the grid must equally assist an electron current moving downwards through that coil ; again, a self-supporting system has been produced.

When considering the effect of the *sustained* action of the anode coil, the effect of the initial action of switching on the H.T. battery is to charge the condenser C. Later, the condenser must start to discharge, and if the oscillatory circuit were disconnected from the grid and filament this discharge would produce a damped oscillatory current ; the energy initially given to the condenser would be dissipated away in the form of heat and æther waves. In such a damped oscillation, the impetus of the electrons flowing out of the condenser charged in one direction is not, by itself, sufficient to give the condenser the same charge in the opposite direction as it originally had ; the maximum P.D. across the condenser becomes progressively smaller as the oscillation dies away.

In the present case, however, the downward flowing electrons through L receive an extra " kick " and, when the direction of electron flow changes, the upward flowing electrons receive a similar kick from the coil in the anode lead The nearer the anode coil is to the grid inductance, the greater is this kick, and it can be made sufficient to compensate for the loss of energy in the form of heat and æther waves, and to charge up the condenser in the opposite direction by an amount equal to its initial charge. It is even possible to over-compensate for that loss of energy ; in which case, the condenser becomes charged to a greater and greater potential in successive swings, until an equilibrium point is reached when the additional energy being dissipated is just equal to the amount which is derived from the additional kick. If the two coils are coupled as shown, but are not wound in the same direction, the direction of the kick will be in anti-phase to the oscillation which it is desired to support, and oscillations will cease.

It will be seen that, as long as the key is pressed, the alternating upward and downward flow of electron currents in L are maintained by energy derived in the form of kicks from the anode coil. In all cases an equilibrium is reached, and an oscillatory current of constant amplitude is maintained. By coupling to an aerial it is possible to radiate an undamped æther wave.

At first sight, the process may look rather like the attempt of a man to lift himself by pulling at his own bootstraps, since the changing currents in the anode coil are caused by the grid becoming alternately positive and negative to filament, and the maintenance of this in constant amount depends in turn on the same changing currents in the anode coil. It must be remembered, however, that it is the deficit of electrons on the anode, due to the battery between anode and filament, which causes any current to flow through the anode coil at all. The grid is not responsible for this current any more than the policeman on point duty is responsible for the presence of traffic in his street. He merely controls its movements along the street, just as the grid controls the flow of current to the anode and therefore through the anode coil. The power which carries the vehicles along is their own engine

power. The power which carries the electrons through the valve and the anode coil, and which ultimately supplies the heat and æther wave power losses in the oscillatory circuit, is derived from the H.T. battery. When the key is released and the battery connection broken, the oscillatory current in the grid circuit dies away because of heat and radiation losses, just as it did in the spark transmitter circuit. The condenser is, of course, not charged again until the key is pressed once more. Thus the oscillation is continuous all the time the key is pressed, dies out as a damped oscillation when the key is released, and nothing more happens until the key is again pressed.

In the circuit shown in Fig. 1, the undamped oscillatory current is not generated in the aerial circuit, but is transferred to it by mutual inductance from the primary oscillatory circuit between grid and filament of the valve. The primary circuit could, however, be utilised as an open oscillatory circuit, and a separate aerial circuit be dispensed with, by substituting an aerial for the primary condenser.

The circuit is sometimes used in heterodyne oscillators.

From another point of view, this simple self-oscillatory system may be regarded as a converter of energy from D.C. to A.C This is a change which can be effected with an efficiency which will be considered later.

An analogy sometimes used in describing the action of a valve oscillator is that of a pendulum clock and its associated spring, or falling weights. The stored energy of the spring, or the potential energy of the weights, is fed at the top of the pendulum in the form of "kicks" by means of the mechanism known as the "escapement." The latter is comparable, in function, to the anode coil of this simple oscillator, the nett effect being to convert energy of one form into the oscillatory energy of motion of the pendulum. The analogy is not quite a perfect one, since the escapement of a pendulum applies the push at the point of maximum potential energy, while in the valve circuit the greatest push occurs close to the point of maximum kinetic energy ; the valve circuit is more nearly equivalent to a pendulum in which the kick is applied at the centre of the swing and not at the ends.

3. **From Reaction Receiver to Oscillator.**—In order that one can better understand the subsequent development of the simple valve oscillator, it is essential to put the above qualitative account into more precise form. It is evident that Fig. 1 represents the basic circuit of a reaction receiver, and it will be recalled that the object of regenerative amplification is to feed back sufficient "negative resistance" (or energy) to neutralise, partially, the damping of the positive resistance in the circuit (Section "D"). It has already been shown that if too much energy is fed back into the tuned circuit, it becomes self-oscillatory and oscillations persist once they are started. Now the basis of a transmitter is an oscillator, and the germ of the latter lies in the reaction receiver.

4. **Phase Relations in the Simple Oscillator ; Self-Oscillatory Criterion.**—The phase relations and quantitative results already obtained with the reaction receiver are here summarised for the sake of convenience. The numbers at the ends of the vectors in Fig. 2 (*b*) represent the order in which the vectors may be studied, or the diagram developed.

Assume an initial oscillatory current of R.M.S. value I begins as soon as the H.T. voltage is switched on when the filament is alight ; this is given by vector I. This current lags about 90° behind the voltage developed between grid and filament, which is the vector V_g and is the *back E.M.F.* developed across the condenser C. This grid input voltage gives rise to an oscillatory anode current I_a which, in practice, is almost in phase with it. Owing to the mutual inductive link between the anode and grid circuits, an E.M.F. will be injected into the latter which will either lead or lag on I_a by exactly 90°, depending upon the sign of the mutual link. The ωMI_a vector is drawn for the self-oscillatory case, the one in which the injected volts are either in phase with the current I that was supposed to start the action, or have a large component in phase with it (*cf.* Appendix "E").

When the anode coil supplies energy in phase with the current I, a self-supporting oscillatory system is obtained ; a **simple vector test or criterion for self-oscillatory conditions has thus been developed.**

Quantitatively the matter is summarised as follows :—

When the anode coil supplies energy in phase with the oscillatory current in the circuit—

Energy introduced into the LC circuit $\doteqdot (\omega M I_a)\, I$ .... in watts.

$\therefore$ Nett loss of energy in LC circuit is at the rate of

$$I^2R - (\omega M I_a)\, I \text{ .... in watts.}$$

$$\doteqdot \quad I^2R - \omega M g_m V_g I \text{ .... since } I_a \doteqdot g_m V_g$$

$$= \quad I^2\left(R - \frac{M g_m}{C}\right) \text{ .... since } V_g = \frac{I}{\omega C}$$

This gives the result that when M is equal to or greater than CR/g_m, oscillations will be continuous or build up respectively.

The result is only approximately true, since it has been assumed that the coil L_2 is small enough for us to neglect its effect of altering the anode volts (and hence I_a) by reason of the voltage drop

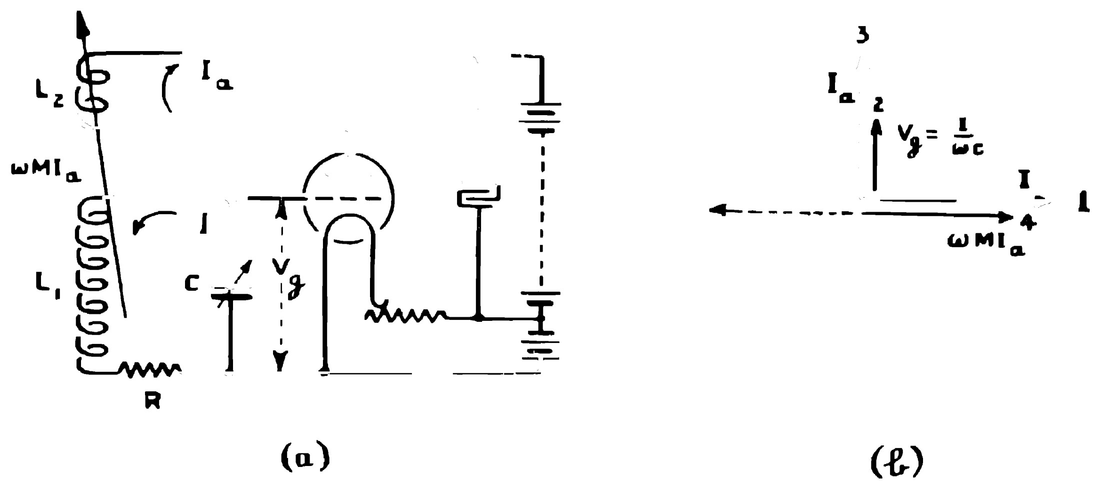


FIG. 2.

across it. This means that the static characteristic has been used, as opposed to the dynamic characteristic which must be used for more accurate work For completely satisfactory results, account must also be taken of the curvature of the dynamic characteristic.

5. **A Simple Oscillator Modified for Higher Power.**—A well-designed circuit of the above type is capable of maintaining quite a large amplitude of oscillatory current, but in practice its use is mainly confined to small lócal oscillators. For transmitters—*i e.*, generators designed to convey electromagnetic energy to considerable distances—a different procedure is usually adopted. In order that more power may be obtained from the oscillatory circuit, the latter is usually connected in the anode lead, and reaction is introduced from the grid circuit ; the positions of the tuned circuit and the reaction coil are simply interchanged.

In general terms, the action can be described in a similar way to that of the circuit of the simple oscillator. If any electric shock is given to the oscillatory circuit by a slight change in current passing through it, an oscillatory current is set up. This oscillatory current, by means of the mutual link between the anode and grid circuits, sets up oscillatory variations of grid-filament voltage at the same frequency. If the sign of the mutual inductance is the correct one, the " injected volts " between grid and filament will provide variations in anode current passing through the oscillatory circuit, and supply power to it so as to tend to maintain the initial oscillation.

In order that the oscillatory action may maintain its initial amplitude, or increase up to a point where limiting conditions supervene, it is not, in itself, sufficient that the sign of $\omega M I_a$ should be

correct ; the coupling must be sufficient to allow the grid-filament voltage variations to supply enough power to the oscillatory circuit to make good the damping losses.

Fig. 3 (*a*) represents the circuit of the modified simple oscillator.

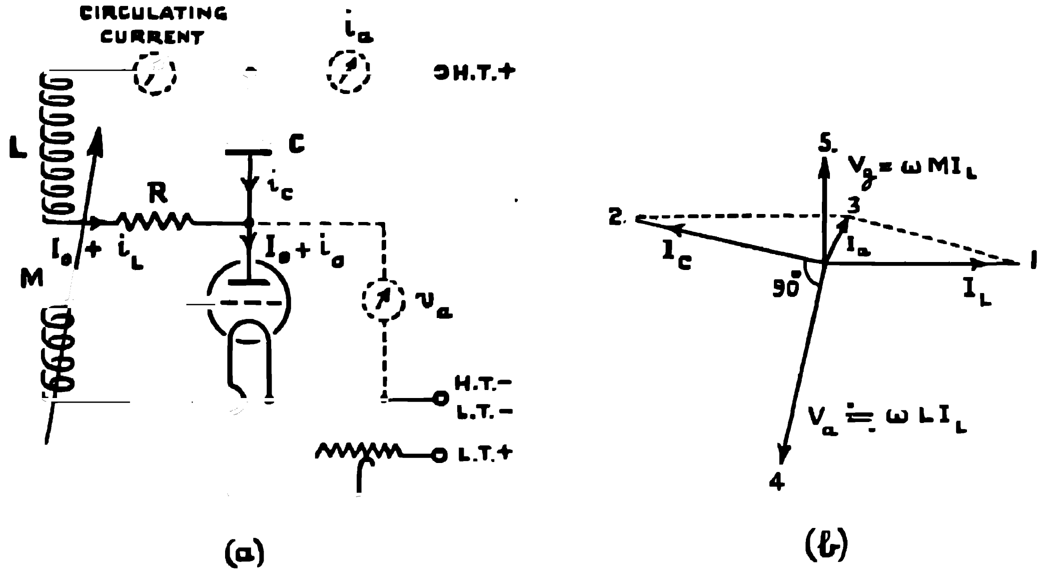


FIG. 3.

If oscillations are set up in the LC circuit, the currents flowing in the different parts of it are as represented in the figure. The total current through the valve is made up of a steady component I_0 plus an oscillating component whose instantaneous value is represented by i_a. The current in the inductance L is similarly composed of I_0 plus an oscillating component i_L ; the current in the capacitive arm of the parallel circuit consists only of an oscillatory component with instantaneous value i_C.

It follows that, vectorially,

$$i_a = i_L + i_C.$$

In practice, I_C is slightly larger than I_L.

6. **Modified Phase Relations ; Oscillator as an Amplifier providing its Own Input.**— It may be observed that, as the variations in current occur at the natural frequency of the LC circuit, which is practically equal to its resonant frequency, the anode circuit may be regarded as almost a rejector circuit to variations of current flowing through the valve, the relative phase relationships of I_L, I_C and I_a being the same as those applicable to the make-up current and the circulating current in a rejector circuit. I_L etc. represent the R.M.S. values of the currents i_L etc. This being the case, the vectors I_L and I_C will be almost 180° apart, and their vector sum will be I_a.

The vector relations are shown in Fig. 3 (*b*), the numbers on the vectors again representing the order in which they may best be considered. Commencing with I_L, I_C is drawn not quite in anti-phase to it. I_a is shown as the vector sum.

There is no oscillatory applied E.M.F. in the circuit (the supply being D.C.), and so the total oscillatory P.D. around the circuit must be zero. Now the oscillatory P.D. (V_R) across the rejector circuit is nearly in phase with the make-up current I_a. In practice, there is not a uniform circulating current, and the current in the capacity arm is slightly greater than that in the inductive arm. In Fig. 3 (*b*) the vector I_C should, therefore, be drawn longer than I_L. Under these practical conditions the make-up current I_a slightly leads V_R, but lags on V_g as shown in the diagram. As the voltage drop across the rejector circuit INCREASES, the anode-filament P.D. (V_a) of the valve must DECREASE ; this means that V_a must always be equal and opposite to V_R, *i.e.*, $V_a = -V_R$.

Hence, V_a is approximately 180° out of phase with I_a; more accurately it may be drawn as a vector 90° ahead of I_C, since it may be regarded as a back E.M.F. across the condenser C, or, alternatively the back E.M.F. across the inductance L. A rough value of this back E.M.F. is given by $\omega L I_L$ the resistance of the inductance being neglected.

With coils wound and connected as shown, and with no current in the grid circuit, the vector diagram represents the instant of time at which i_L is zero but approaching positive values, and increasing towards its maximum value. i_L acts away from the ***common point,*** provided by the filament (Vol. 1 paragraph 161), and induces an E.M.F of value $M\frac{di}{dt}$ in the grid coil; this also acts away from the common point and makes the grid positive with relation to it. As time goes on, $M\frac{di}{dt}$ will become smaller as i_L approaches its maximum positive value, since the rate of change of current becomes smaller; the vector V_g (or $\omega M I_L$) must therefore be drawn *leading* I by 90°, so making V_g very nearly in phase with I_a (*cf.* Appendix "E".)

The circuit is self-oscillatory because the injected volts V_g have a large component in phase with I_a, and power is supplied tending to maintain the initial oscillation. The power supplied is given by $V_g I_a$ Cos ϕ, where ϕ is the small angle between V_g and I_a; for practical purposes Cos ϕ may be taken as unity (paragraph 10).

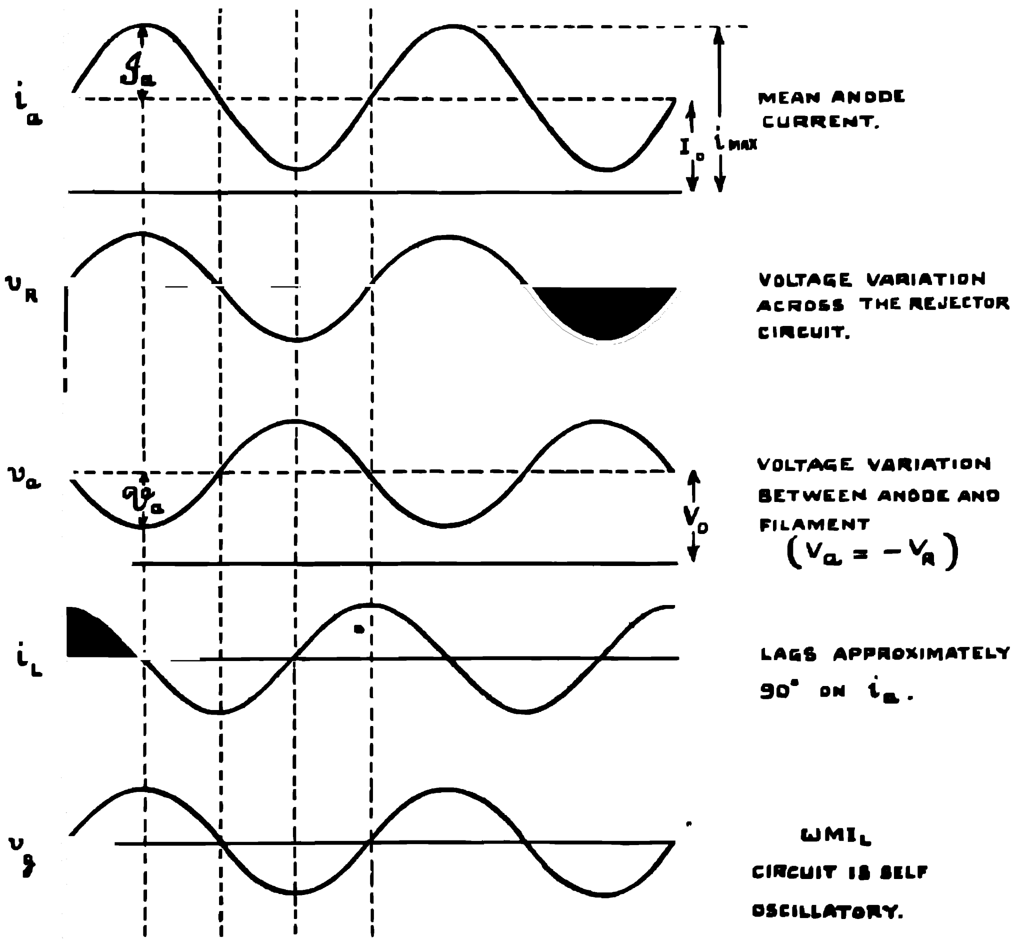


FIG 4

Erratum—Fig. 4. Re-draw curve for i_L to lag 90° on i_a; i.e, 180° out of phase with the curve for i_L shown, which should be deleted.

Fig. 4 represents these phase relationships in the form of curves. This method of representation has the advantage of showing more clearly the approximate anti-phase relationship between I_a and V_a. As i_a increases, v_a decreases, and this is the result which is sometimes considered as the application of Lenz's law to a valve circuit.

For systems of this kind, it is sometimes stated as a CRITERION DETERMINING SELF-OSCILLATORY CONDITIONS that V_a should be approximately 180° out of phase with V_g. In considering the further development of the valve oscillator, a criterion in this form will be particularly useful.

It will be noted from the curves that when v_g has its maximum positive value, v_a has its minimum value. Now the oscillator will not function when the grid is more positive than the anode, and this is one of the factors that, in practice, limits the value of V_g.

The appearance of the circuit of Fig. 3 (*a*) suggests an amplifier with a "tuned anode" output impedance. Now the parallel is a very complete one, and for many purposes it is most convenient to regard an oscillator of this kind as being equivalent to an amplifier which provides its own input.

It is possible to construct a working model oscillator of this type, in which frequencies of the order of 1 cycle per second can be produced. Instruments may be inserted in the circuit in positions shown in Fig. 3 (*a*), and it is very instructive to watch the movements of the pointers. In such a model it is very essential to use instruments which are practically dead-beat, and which are similar in all respects other than that of the coil-circuit resistance. Under these conditions, the phase difference between the pointers and the oscillatory quantities which they represent will not be a serious matter.

Finally, it may be noted that the slightly out of phase relationship between V_g and I_a produces a dynamic characteristic which is elliptical in shape. For many purposes, to simplify treatments, this ellipticity is here ignored, and the dynamic characteristic is treated as a straight line.

7. **Further Development of the Modified Oscillator ; The Hartley Circuit.**—The above statement of the criterion determining self-oscillatory conditions, is in a form that stresses the optimum conditions. In point of fact, generation of self-oscillations will usually be possible when the oscillatory grid voltage V_g is more than 90° out of phase with the anode oscillatory voltage V_a, the best conditions being when the two are in anti-phase. In addition to the energy supplied to the oscillatory circuit being in the correct phase, it must also be sufficient in amount to overcome the damping losses.

A great variety of circuits can be arranged which satisfy the conditions for self-oscillation. It is out of the question to deal individually with even a small fraction of the possible valve transmitting circuits that may be encountered in practice. In the present instance it is only proposed to consider briefly two simple developments of the modified oscillator, both of which have particular reference to Service practice. A further number of circuits are considered later, when an attempt is made to classify valve transmitters.

Fig. 5 (*a*) and (*b*) represents two forms of the same type of self-oscillatory circuit, usually known as a Hartley circuit. Fig. 5 (*b*) is the simpler of the two forms and is called a "series fed" circuit, since the valve, the high tension battery, and the tuned circuit, are all in series. Fig. 5 (*a*) is another form of series fed Hartley circuit.

The action may be explained as follows :—The oscillation set up in the tuned circuit produces an oscillatory voltage across the portion of the inductance from G to F, *i.e.*, between grid and filament. The optimum condition for self-oscillation is that this voltage should be in anti-phase to the anode-filament oscillatory P.D., *i.e.*, the P.D. across the portion AF of the tuned circuit inductance. This is secured by connecting the filament to a point on the inductance, intermediate between the points connected to anode and grid. The point F is obviously always at an intermediate potential between the potentials of A and G. If A is positive to F at any instant, G must be negative to F at the same instant, and so the P.Ds. from A to F and G to F are always in opposite directions. These are respectively the anode-

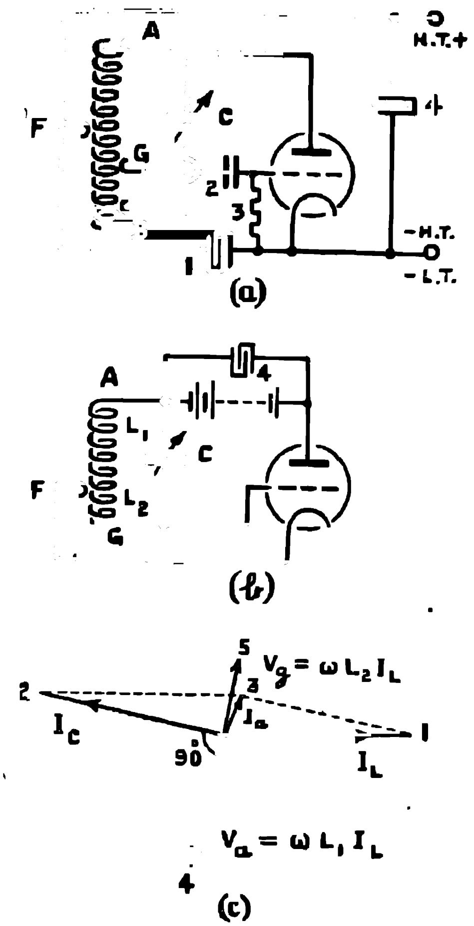


FIG. 5.

filament and grid-filament P.Ds., and so V_a and V_g are in anti-phase and the circuit self-oscillatory.

In both forms of the circuit, the "filament tap" F provides a means of adjusting the magnitude of V_g. In Fig. 5 (*a*), this adjustment may also be performed by means of the "grid tap" G, assuming the tapping at F to be fixed. Also in Fig. 5 (*a*), the H.T. supply to the anode is fed through the tap F, which is accordingly called the "anode tapping point," the adjustment of the anode tap being further considered below. The large condenser (1) is necessary in order to avoid short-circuiting the H T. supply between grid and filament, and to provide an R/F earth between F and the filament, in order that these two points may always be at the same oscillatory potential.

The function of the grid condenser and leak (2) and (3) is two-fold; it insulates the grid from the high tension supply, and also produces an automatic negative bias which increases the efficiency of the oscillator This latter point is treated in detail later on. The big condenser (4) is an ordinary R/F by-pass condenser across the H.T. battery.

The vector diagram representing the operation of an oscillator of this kind is shown in Fig. 5 (*c*). It will be seen to be essentially the same as that of Fig. 3 (*b*), with the difference that V_g and V_a are exactly in anti-phase in this case. Whether or not the vector V_g is more nearly in phase with I_a, will depend on other considerations, involving the relative magnitudes of I_C and I_L.

For the reasons given in paragraph 6, the vector I_a will usually lead V_R by a small amount.

8. **Evolution of the "Divided Circuit."**—The Hartley circuit is a very simple form of an "amplifier that provides its own input." There are many varieties of it, and some of them present

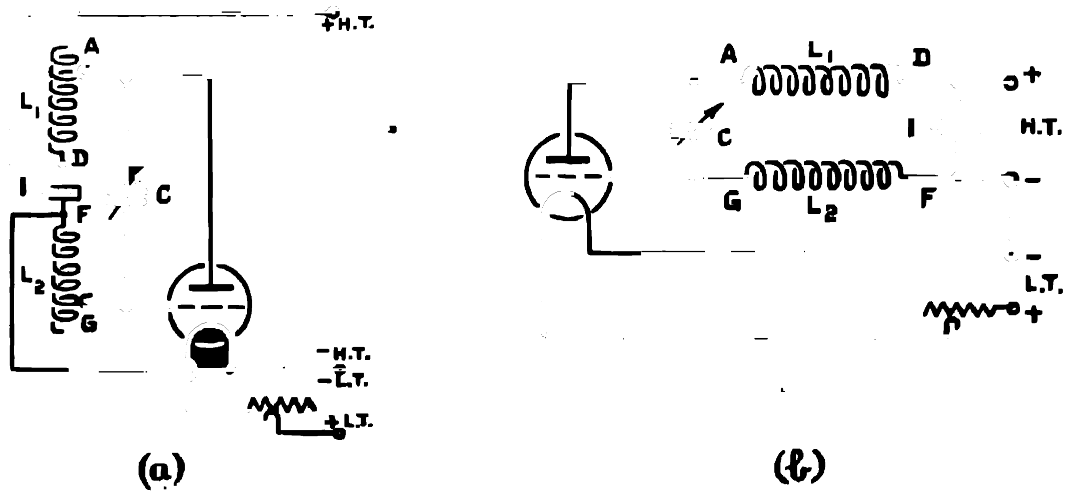


FIG. 6.

a very different appearance. One of these with application in the Service, is known as the " divided circuit," and its evolution may be treated in two stages, shown diagrammatically in Fig. 6 (*a*) and (*b*).

Fig. 6 (*a*) is another form of series fed Hartley circuit. It will be seen that the inductance is in two portions, separated by a large condenser (1) ; this condenser is large in comparison with condenser C—with which it is in series—and hence exercises very little control over the frequency of the LC circuit. The H.T. positive terminal is connected to the anode side of this condenser, and the other plate is connected to the filament. The reactance of this condenser is negligible at radio-frequencies, but it effectively isolates the H.T. positive terminal from the grid, and also prevents the short-circuiting of the H.T. supply. The position of this large condenser in the inductance of the tuned circuit also provides a method of varying the amount of " grid excitation," *i.e.*, the oscillatory voltage between G and F, to obtain the greatest efficiency.

The circuit of Fig. 6 (*a*) may be re-drawn as in (*b*), and it is from the appearance of this diagram that the name " divided circuit " arises

9. **The Colpitts Circuit.**—Fig 7 represents an oscillator of the same nature as that of Fig. 5. In this case the filament tap is taken from a point in the capacitive arm of the tuned circuit, the latter being split into C_1 and C_2 as shown. The explanation of the action of the circuit is precisely the same as that of the Hartley oscillator ; in this case, the grid excitation might be described as " direct capacitive," the Hartley oscillator having " direct inductive " grid excitation. When an oscillatory voltage is developed across the tuned circuit, the filament potential is intermediate between the grid and anode potentials, V_a and V_g are in anti-phase and the circuit is self-oscillatory.

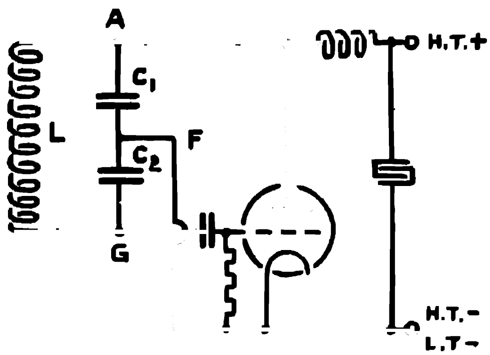


FIG. 7.

The circuit of Fig. 7 will be suitable for low and intermediate frequencies ; it will be seen later that at high frequencies it is often quite sufficient to place the tuned circuit between the anode and grid, and to rely on the inter-electrode capacity of the valve to provide the necessary grid excitation using the Colpitts principle.

10. **Power in the Oscillatory Circuit.**—If an oscillator of the type of (say) Fig. 3 is rendered non self-oscillatory by (say) breaking a lead in the grid circuit, the anode current and anode-filament P.D. will remain steady at values denoted by I_0 and V_0, as shown in Fig. 4. The power being dissipated at the anode of the valve is given by the product $I_0\ V_0$.

Under self-oscillatory conditions, the peak value $\mathcal{I}_a$ of the oscillatory anode current is clearly limited by the magnitude of I_0 ; $\mathcal{V}_a$, the oscillatory anode-filament P.D., is similarly limited by the magnitude of V_0, the voltage of the H.T. battery. It would therefore appear that the larger the value of the mean anode current and the D.C. voltage applied to the anode, the larger may be the power isolated in the **oscillatory circuit**, now often called a " tank circuit." Hence the use of thousands of volts on the anodes of transmitting valves.

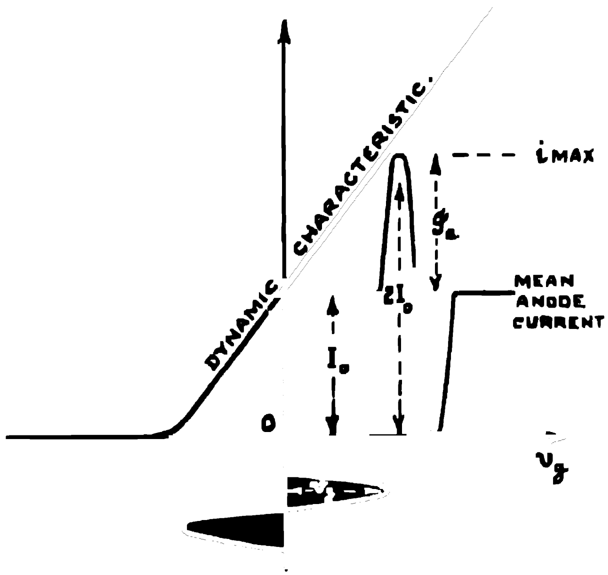
The oscillatory power is given by the product of the R.M.S. anode current and the R.M.S. voltage, assuming the " power factor " (Cos ϕ) equal to unity, *i.e.*, I_a in anti-phase to V_a :—

$$\text{Oscillatory Power} = I_a\ V_a = \frac{\mathcal{I}_a}{\sqrt{2}}\,\frac{\mathcal{V}_a}{\sqrt{2}} = \frac{\mathcal{I}_a\mathcal{V}_a}{2}$$

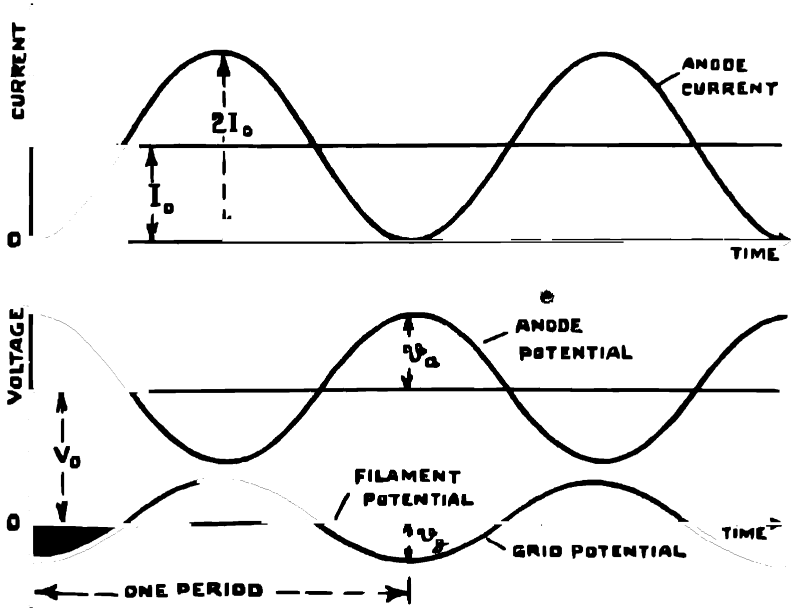
The R.M.S. value of the " circulating current " in the LC circuit is given by

$I^2 R = I_a V_a$............ (R is the resistance of the LC circuit)

11. **Class " A " Operation ; Theoretical Maximum Efficiency.**—If the working point of an oscillating valve is such that oscillations occur within the straight part of the dynamic characteristic, in modern terminology the valve is said to be working under Class " A " conditions. It is analogous to a " mid point " biased amplifier and is represented diagrammatically in Fig. 8.



(a)



(b)

FIG. 8.

If we assume that suitable adjustments are made to the oscillator so that $\mathcal{I}_a$ and $\mathcal{V}_a$ both have the maximum theoretical values of I_0 and V_0, we then have :—

$$\text{A.C. output power} = \frac{I_0 V_0}{2}$$

$$\text{D.C. input power} = I_0 V_0$$

since in Class " A," the mean values are unaltered during oscillatory conditions.

Hence, the efficiency of power conversion is given by :—

$$\eta = \frac{I_0 V_0/2}{I_0 V_0} = 50 \text{ per cent.}$$

The meaning of this " ideal " result is, that the power dissipation is equally divided between that expended in heating the anode of the valve and that available in the LC circuit. It should be noted that if oscillations suddenly ceased, for any reason, almost all of the available power $I_0 V_0$ would then have to be dissipated at the anode ; the valve will be operating under static conditions, the A.C. resistance being bigger than under dynamic conditions and constituting the main impedance of the circuit. If a transmitting valve is called upon to dissipate the whole of the power taken from the H.T. supply instead of a percentage of it (say, between 30 and 50 per cent.), the load may exceed the anode rating of the valve and the valve may be burnt out.

The power taken from the H.T. battery is the same under oscillatory and non-oscillatory conditions, so that, more generally, the power dissipated at the anode in the form of heat is given by :—

$$\text{Anode dissipation} = I_0 V_0 - \frac{\mathcal{I}_a \mathcal{V}_a}{2}$$

The above result can be shown mathematically—though no more clearly—by expressing the power dissipation at the anode in the form of $(I_0 + \mathcal{I}_a \sin \omega t)\ [V_0 + \mathcal{V}_a \sin (\omega t + \pi)]$, and subsequently simplifying. The oscillatory terms which appear in the simplification have a mean value which is zero, if taken over a whole cycle.

The 50 per cent. efficiency result may also be compared with the adjustment of an amplifier to give ***maximum power ouput*** ; in that case it has been seen that the output impedance must be designed to be equal to the A C. resistance of the valve, and that implies equal expenditure of power within the valve and outside it.

In this connection, it is also of interest to recall that to obtain ***maximum undistorted ouput*** from an amplifier under Class " A " conditions, it is necessary that the output impedance should be at least twice the A C. resistance of the valve. This result was obtained by arranging to work the valve under conditions which did not involve the flow of grid current. Applied to the case of a Class " A " oscillator, this would give an efficiency of only 25 per cent.

In practice, an efficiency of 50 per cent. is quite unattainable, under these conditions, for various reasons. It is due partly to the dissipation of energy by grid current, and very largely to the fact that $\mathcal{V}_a$ can never be equal to V_0.

The anode-filament P.D. must never be less than the maximum positive grid-filament P.D. ; otherwise the grid robs the anode of current, and instead of the anode current rising to a maximum at the point of minimum anode-filament P.D., it will fall to a very low value. This produces unwanted harmonics and is detrimental to the efficiency. Further, the heavy grid current may lead to a large secondary emission from the grid. When the number of secondary electrons emitted by the grid, due to grid electron bombardment, exceeds the primary electrons bombarding it, the nett grid current will flow in the reverse direction ; this reverse grid current, flowing through the grid-leak resistance, raises the grid to a positive potential which tends to become larger with great rapidity. The circuit becomes heavily damped and the amplitude of oscillations rapidly decreases. Finally, when static conditions supervene, the anode is always positive to the grid but the large grid secondary emission then causes a rush of electrons to the anode, resulting in a greater production of heat at that electrode than the valve is constructed to dissipate, and hence damaging the valve. This phenomenon is called **blocking** (K.47).

Since the minimum anode-filament P.D. occurs at the same instant in a cycle as maximum positive grid-filament P.D., it follows that the oscillatory anode-filament voltage—which is $\mathcal{V}_a$—must be less than the voltage of the H T. supply—which is V_0—by at least the amount of the grid voltage variation.

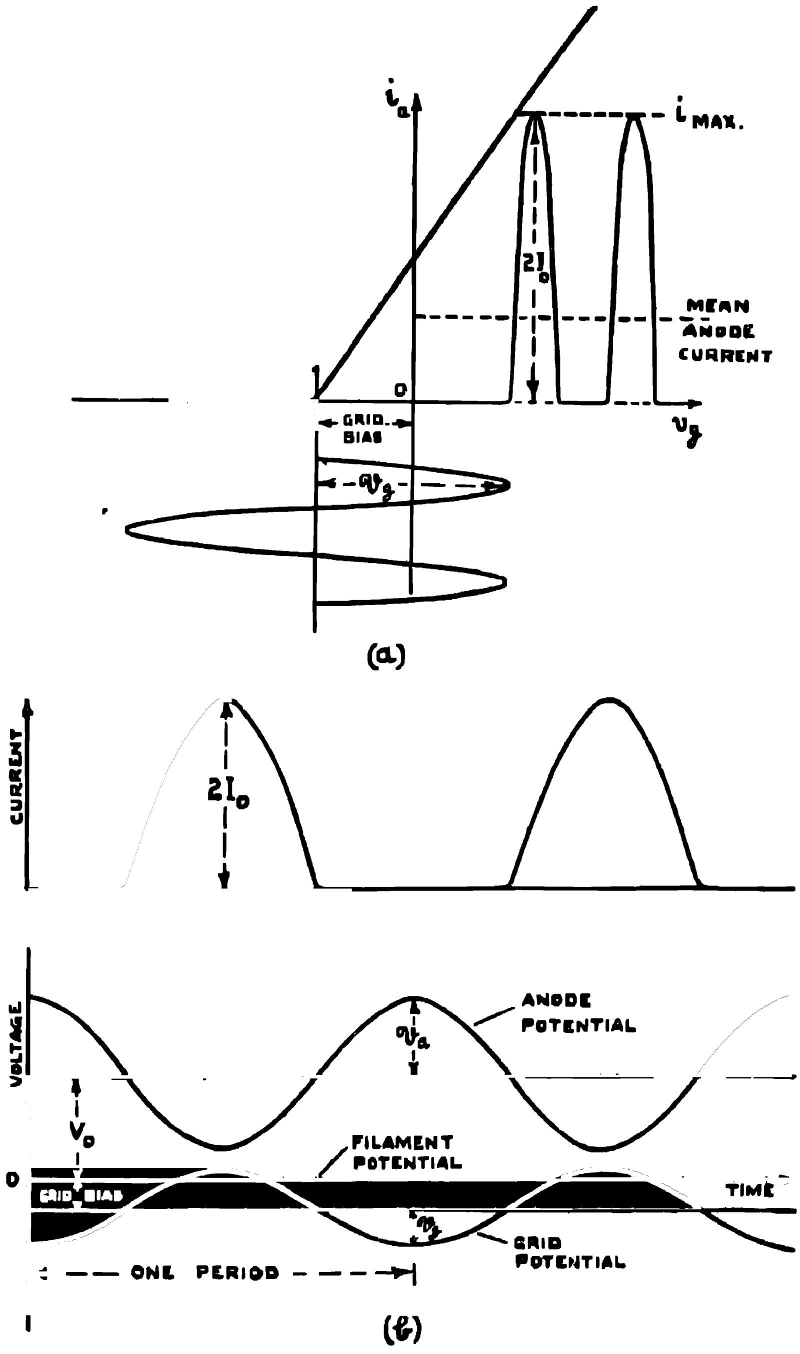


FIG 9.

12. **Class "B" Operation; Efficiency with Working Point at the Lower Bend.**—Although Class "A" operation gives an anode current wave form which is approximately sinoidal in shape, or a faithful copy of the grid input wave form, a power conversion efficiency of less than 50 per cent. can hardly be accepted. Efficiencies up to about 70 per cent. can be obtained if the working point of the valve is adjusted to be at about the middle of the lower bend, approximately as in Fig. 9. This is a definition of Class "B" operation. Anode current only flows during the positive half cycles, and its mean value is less than under corresponding Class "A" conditions; in the non-oscillatory state it is almost zero.

In producing Fig. 9, it has been assumed that the valve is the same as that employed in Fig. 8; the grid swing has been approximately doubled in order that the maximum anode current $i_{max.}$ should be the same in the two cases.

In the Class "B" case, the anode current wave form is no longer sinoidal, but it can be shown to have a FIRST HARMONIC CONTENT OF ABOUT THE SAME AMPLITUDE as that in the corresponding Class "A" case. It therefore follows that, with suitable adjustment of the anode impedance, the same fundamental, or first harmonic, oscillatory power output may be obtained as in Class "A" operation, the D.C. input being relatively smaller. Of two devices, that which gives the same output for lower input is the more efficient.

In greater detail the above matter may be explained with the help of the curves of Fig. 10.

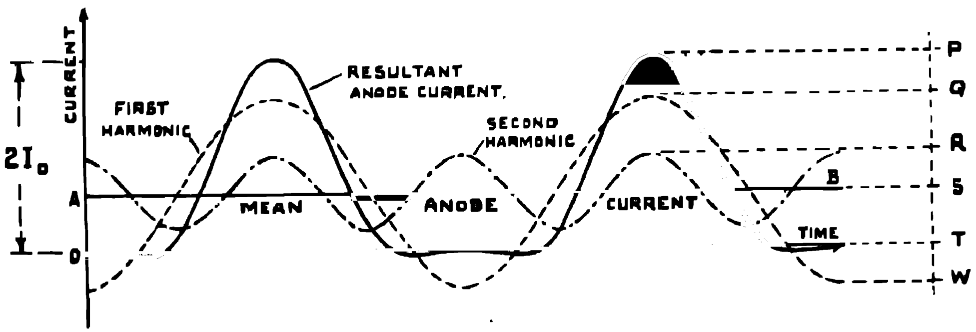


FIG. 10.

A first harmonic current wave is shown, together with a second harmonic, *i.e.*, a current at twice the frequency of the first harmonic. The amplitude of the second harmonic is taken as a third of that of the fundamental, and at t = o it lags on the latter by an eighth of the fundamental period. They are plotted about a common axis AB, and the resultant wave form, which is the sum of these two components, is indicated by the full line curve. Mathematically expressed, the resultant curve is the sum of the two terms in an expression of the form—

$$y = \sin t + \frac{1}{3} \sin\left(2t - \frac{2\pi}{8}\right).$$

It will be observed that the resultant curve is very similar to the anode current wave form of Fig. 9, and for approximate work it will be regarded as identically the same, the zero current line—or X-axis of the graph—being shifted down by an amount OA. The new mean value of the anode current is therefore OA, and its magnitude is less than I_0—since it is less than $\mathcal{I}_a$—the mean value under 50 per cent. efficiency conditions. From this it is clear that the D.C. power input is smaller.

In regard to the amplitude of the first harmonic content, we can see from the figure that

$$\begin{aligned} PT &= QS + PQ + (SW - TW) \\ &= QS + RS + (QS - RS) \\ &= 2QS. \end{aligned}$$

This implies that the first harmonic oscillatory output is the same under the two sets of operating conditions; the same output is obtained with a lower input and the arrangement is relatively more efficient.

★13. NUMERICAL ESTIMATE OF EFFICIENCY WITH CLASS " B " OPERATION. Some idea of the numerical value of the efficiency under Class " B " conditions may be gleaned as follows, making the assumption that the anode current wave form may be taken as sinoidal during the positive half cycles of grid voltage, and zero during the negative halves.

With reference to Fig. 9, the mean value of a sine wave ($i_{max.}$ sin ωt) over one HALF CYCLE is given by—

$$\frac{\omega}{\pi}\int_0^{\frac{\pi}{\omega}} i_{max}\sin\omega t.\,dt = \frac{2\,i_{max.}}{\pi}$$

But during the next half cycle the mean value is nil. Hence, the mean anode current in Class " B " working—

$$= \frac{i_{max.}}{\pi} = \frac{2\,I_0}{\pi} \quad \ldots\ldots\ldots\ldots\ldots \qquad \text{(Ref. Fig. 8)}$$

$$\doteq \frac{2}{3}\text{(mean anode current in Class " A " working).}$$

The D.C. power input giving the same oscillatory output is therefore relatively less, and, assuming 50 per cent. efficiency for Class " A "—

$$\text{Efficiency using Class " B "} = 50 \times \frac{\pi}{2} = 78\cdot5 \text{ per cent.}$$

This is an efficiency not fully realised in practice.

14. Class " C " Operation ; Working Point beyond the Lower Bend.—In the case of Class " B " operation, it should be observed that although the anode current wave form is not sinoidal in shape, the positive half cycles may be regarded as faithful copies of the grid-filament oscillatory input: The Class " B " amplifier gives approximately linear amplification and a modulated wave form is not distorted.

Still higher power conversion efficiencies can be produced by operating a valve with more than sufficient grid bias to reduce the anode current to zero when there is no oscillatory input between grid and filament. The working point will then be considerably beyond the " cut off " point and may correspond to a bias equal to as much as twice the cut off bias. This is termed Class " C " operation, and efficiencies of the order of 85 per cent. are sometimes possible.

Anode current no longer flows for the whole of the positive half cycles, which are therefore no longer even moderately faithful copies of the grid oscillatory input. In order to obtain peak values of anode current equal to those obtained under corresponding Class " A " or Class " B " conditions, the grid oscillatory input must be considerably bigger.

Figs. 8 and 9 emphasise the increase in grid swing that marks the difference between Class " B " and Class " A " operation ; with the working point beyond the cut off, in Class " C " operation, the positive " flicks " of grid voltage will need to be very high in order to produce peak values of anode current equal to those obtained by the other methods of operation. The name " flick impulsing " is sometimes applied to Class " C " operation. As in the case of Class " B," many harmonics are present in the anode current wave form, and, unless steps are taken, this may lead to the radiation of energy at undesirable frequencies.

15. Automatic Bias for Class " B " and Class " C " Operation.—The requisite negative bias for the adjustment of the working point to the lower bend, or beyond it, may be obtained automatically by the introduction of a condenser and leak into the grid circuit. (*Cf.* N. 41).

When the valve is generating continuous oscillations, the grid oscillatory potential varies through large values on either side of its steady value, and, when it is positive to the filament, grid current flows. The condenser thus acquires pulses of negative charge, and with a high resistance leak this is slowly drawn away, with the result that the mean grid potential, relative to the filament, is held at a negative value equal to the IR drop across the resistance. (This may be compared with the steady negative grid bias attained by a grid current detector when receiving C.W.). The more powerful the oscillations the greater is the negative potential of the grid, and by designing the leak suitably, the grid can always be kept at any desired value of negative potential.

The values of the condenser and leak resistance are important. If the time constant CR of the combination is too large, the grid builds up so large a negative potential that the amount of power transferred from the valve to the oscillatory circuit is insufficient to maintain oscillations. When oscillations cease, no more grid current flows, and the condenser discharges through the leak until reaching the point at which the potential of the grid is such that oscillations again become possible. Under these conditions, oscillations will be continually starting and stopping, a phenomenon sometimes known as "squegging" or "grid tick." This is obviously a thing to be avoided when continuous oscillations are desired; the remedy is to reduce the CR value of the condenser and grid leak combination, and it will usually be better to reduce the value of the grid leak resistance.

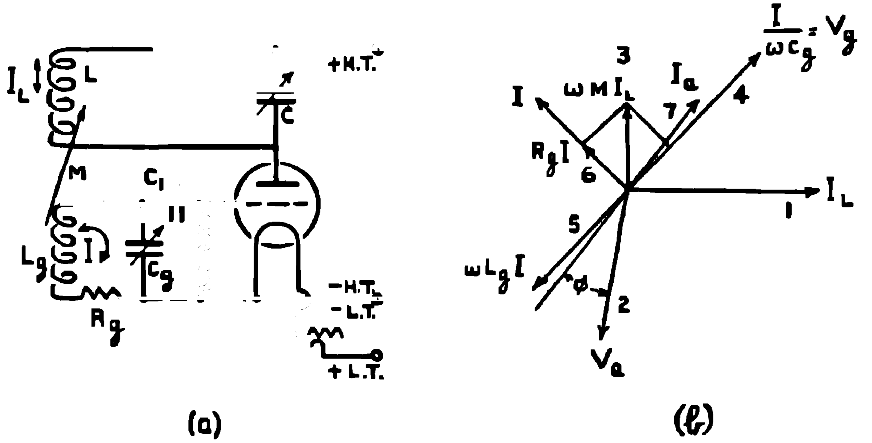
An unnecessary increase in the proportion of harmonics in the anode current wave form will result if the variation of mean grid potential during the radio-frequency cycle is at all pronounced. This occurs if the condenser has too small a capacity. The charges which it acquires through the grid current flowing during the positive half cycles, followed by the losses through the leak resistance during the negative half cycles, produce corresponding variations of grid potential, which are larger the smaller the capacity of the condenser. Hence the condenser must be large enough to obviate pronounced fluctuations in the mean negative potential of the grid. (*Cf.* H. 5).

In the case of oscillators, automatic biasing is almost always used; if negative bias were to be applied by a battery, and the working point adjusted to be beyond the cut off, no energy would be fed to the oscillatory circuit and oscillations would never start.

16. **Provision of High Grid Excitation; the Tuned Grid Circuit.**—It has been observed that the requisite grid exciting voltage must be of considerable magnitude in the case of Class "B" and Class "C" operation. In general, the way in which this is obtained will depend upon the nature of the self-oscillatory circuit under consideration.

Considering the particular case of an oscillator of the type shown in Fig. 3, the grid excitation becomes larger if the coupling is tighter. It can also be made greater in another simple way.

With only an inductance in the grid circuit, the grid voltage is given by $\omega M \times$ current in the inductive branch of the oscillatory circuit. This voltage between grid and filament can be increased very much if a condenser is added across the inductance—as in Fig. 11 (*a*)—and this parallel circuit tuned to resonance.



(a) (b)

FIG. 11.

Under these conditions, a voltage ωMI_L induced into this circuit at its resonant frequency sets up a circulating current round it given by $I = \frac{\omega MI_L}{R}$, where R_g is the resistance of the circuit.

The actual voltage applied between grid and filament is that across the inductance, or the capacity, and is $\frac{I}{\omega C_g}$, or $\omega MI_L \times \frac{1}{\omega C_g R_g}$. This is greater than ωMI_L if $\frac{1}{\omega C_g}$, the capacitive reactance, is greater than R_g, the resistance, which is certainly the case.

The theory given above is the same as that for a tuned secondary circuit coupled to the aerial in a receiving circuit, in which the voltage across the condenser is greater than that introduced into the circuit.

The question of phase relationships, however, complicates matters in this case.

The current I is in phase with the voltage applied to the circuit, and so the voltage between grid and filament is now 90° OUT OF PHASE with the induced voltage ωMI_L, and therefore in phase or anti-phase with I_L itself.

Thus the vector diagram of Fig. 3 (*b*) would not hold good. V_g, as we saw, had to be 90° out of phase with I_L, for I_a to be 180° out of phase with V_a, and hence to allow a supply of energy to be given to the oscillatory circuit. Therefore, if the grid circuit is tuned to the same LC value as the anode oscillatory circuit, although the resultant voltage applied to the grid is greatly increased, yet the phasing is—in general—incorrect for the maintenance of oscillations.

It is quite easy to show that if the phase angle between I_a and V_a is $(180° - \phi)$, the power available for the LC circuit is $I_a V_a \cos \phi$. In the case above, $\phi = 90°$, and the power is nil.

THE ACTUAL PROCEDURE EMPLOYED, THEREFORE, IS TO TUNE THE GRID CIRCUIT ONLY PARTIALLY, an intermediate step between leaving it resonant and leaving it aperiodic.

In effect, the injected voltage ωMI_L ceases to be the grid exciting voltage V_g, and becomes the applied voltage in a non-resonant series circuit. As in other cases, the numbers on the vectors in Fig. 11 (*b*) represent the order in which they may conveniently be studied.

Under these conditions, the big grid voltage changes required when a large LC value is used in the oscillatory circuit (with correspondingly small ω) are easily achieved through the grid circuit being partially resonant, while the angle ϕ of departure from the anti-phase relationship of I_a and V_a is intermediate between zero and 90°, so that power is supplied to the LC circuit. In other words, I_a is increased by this arrangement to such an extent that the power supplied $I_a V_a \cos \phi$ is increased, although the power factor, $\cos \phi$, has decreased from its maximum value of unity since I and V_a are not now approximately in anti-phase.

The resonant frequency of the grid circuit is always adjusted to be above that of the oscillatory circuit, *i.e.*, C_g is kept relatively small in order that $\frac{I}{\omega C_g}$ may be large.

As a safety measure, resistance is introduced into the grid circuit by means of lamps, to limit the value of the current in case of accidental tuning to resonance during adjustment. Without this precaution, the current at resonance would damage the grid coil.

Fig. 11 also shows the grid leak and condenser, the function of which is the automatic provision of the necessary bias for Class "B" or Class "C" operation.

17. **Maximum Power Conditions ; Saturation Current.**—In paragraph 10 it was shown that the power in the oscillatory circuit is given by $\frac{\mathcal{I}_a \mathcal{V}_a}{2}$, and in paragraph 11 it was demonstrated that $\mathcal{V}_a$ must be less than the voltage of the H.T. supply V_0 by at least the amount of the grid voltage variation.

If sufficient energy in the right phase is supplied to an oscillatory circuit, oscillations will build up (*cf.* paragraph 4) until some limiting condition supervenes. In Class "A" operation, the working point may be anywhere on the straight part of the dynamic curve.

If it is somewhere between the mid point and the lower bend, it can be shown that oscillations will be built up until the peaks of anode current reach the lower bend, that is, until $\mathcal{I}_a \doteqdot I_0$. If the coupling is sufficient to cause further increase in amplitude, a limiting point is reached because the slope of the dynamic curve becomes effectively less as the excursion of grid voltage past the

bend increases. Finally, an amplitude is reached at which the inequality between M and CR/g_m becomes an equality, and oscillations are maintained at this constant amplitude (*cf.* paragraph 4; a more accurate treatment using the slope of the dynamic characteristic g'_m is given in paragraph 19.).

If the working point is moved up the dynamic curve, the amplitude of the oscillatory anode current continues to adapt itself to the curve, and $\mathcal{J}_a$ grows bigger and bigger until reaching the point exactly midway between the upper and lower bends. At that point the oscillations may be said to fill the valve, the positive and negative peaks extending to the upper and lower bends respectively. Further increase of the maximum anode current $i_{max.}$ is not possible because of the saturation conditions. Hence, the oscillatory anode current has its maximum value when the working point is at the middle of the dynamic curve, its magnitude being equal to half the value of the saturation current ; in symbols we have—

$$i_{max.} = I_s \quad \text{and} \quad \mathcal{J}_a = \frac{I_s}{2}.$$

This is to be regarded as a special case of Class " A " operation, and is represented diagrammatically in Fig. 12

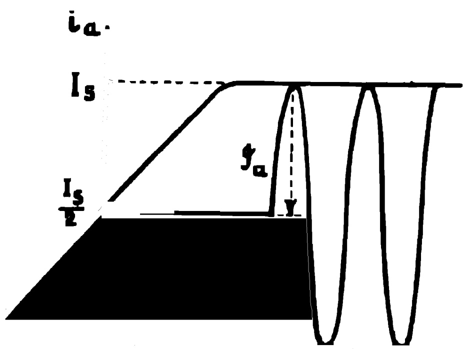


FIG. 12.

Under these conditions, if the anode impedance is properly matched to the valve, maximum power will be developed in the oscillatory circuit.

In order to develop maximum power in the case of Class " B " and Class " C " operation, it is necessary that the power impulse given to the circuit, during the " flicks " of positive grid volts, should be sufficient to traverse the whole of the dynamic characteristic between the bends.

18. **Anode Tapping Point ; Impedance Matching.**—It has been observed (paragraph 11), that in order to supply power with MAXIMUM EFFICIENCY under sinoidal (or Class " A ") conditions, the oscillatory anode current $\mathcal{J}_a$ and oscillatory anode volts $\mathcal{V}_a$ must have the maximum values which are possible under the given operating conditions. As with an ordinary amplifier, this adjustment can be achieved by matching the anode output impedance to that of the valve, the oscillator being regarded as an amplifier in which the input comes from the apparatus itself, instead of from an outside source.

The method of effecting this impedance matching will depend on the nature of the self-oscillatory circuit, and on the frequency in use. It is proposed to consider in detail the particular case of the circuit in Fig. 3.

Now $\mathcal{J}_a$ and $\mathcal{V}_a$ are interdependent (equation (1) below) and—with reference to Fig. 4—their values are determined by the following relations :—

$$\mathcal{J}_a \doteqdot I_0 \quad \text{and} \quad \mathcal{V}_a \doteqdot V_0.$$

But

$$_{a} = \mathcal{J}_a\left(\frac{L}{CR}\right) \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (1)$$

where L/CR is the impedance of the parallel circuit at resonance.

Hence

$$\frac{L}{C} = \frac{R\mathcal{V}_a}{\mathcal{J}_a} = \frac{RV_0}{I_0}$$

(With mid point biasing, $I_0 = \frac{I_s}{2}$).

The product of L and C (the LC value of the oscillatory circuit) is determined by the frequency it is desired to transmit. In addition, the result just obtained shows that a definite ratio of L to C is necessary if power is to be developed in the oscillatory circuit under conditions of maximum efficiency. The practical problem is, therefore, to be able to vary the ratio of inductance to capacity without affecting the tuning of the circuit. This is solved by including only part of the inductance in one side of the parallel circuit, and the remainder in series with the capacity in the other side, as shown in Fig. 13.

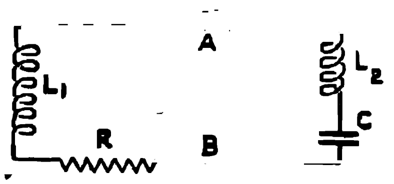


FIG. 13.

If this is done, the natural frequency of the circuit remains unaltered, provided the damping losses are negligible, but the combination of L_2 and C acts effectively as a condenser of increased value, *i.e.*, the total reactance of L_2 and C is series is capacitive, but less than that of C alone.

If this equivalent capacity is written C_1, then

$$\frac{1}{\omega C_1} = \frac{1}{\omega C} - \omega L_2 = \frac{1 - \omega^2 L_2 C}{\omega C}.$$

The natural frequency of the circuit is given by $f = \frac{\omega}{2\pi}$, where

$$\omega^2 = \frac{1}{L_1 C_1} = \frac{1 - \omega^2 L_2 C}{L_1 C}, \quad \ldots\ldots\ldots\ldots \quad (2)$$

i.e., $$\omega^2 (L_1 C + L_2 C) = 1, \text{ or } \omega^2 = \frac{1}{(L_1 + L_2)\, C} = \frac{1}{LC} \quad \ldots\ldots\ldots\ldots \quad (3)$$

which shows that the natural frequency is unaltered.

Also $$C_1 = \frac{1}{\omega^2 L_1} = \frac{LC}{L_1}\ . \quad \ldots\ldots\ldots \left[\text{from (2) and (3).}\right]$$

The ratio of inductance to capacity in this circuit is therefore $\frac{L_1}{C_1} = \frac{L_1^2}{LC}$, and the effective resistance of the circuit is $\frac{L_1}{C_1 R} = \frac{L_1^2}{LCR}$. By this device the effective inductance of the circuit has been decreased, the effective capacity increased, and the effective resistance decreased.

Equating this effective resistance to the A.C. resistance of the valve--

$$\frac{L_1^2}{LCR} = r_a$$

or $$\omega^2 L_1^2 = Rr_a \text{ and hence } \omega L_1 = \sqrt{Rr_a}$$

$$\therefore \quad L_1 = \frac{\sqrt{Rr_a}}{\omega} = \sqrt{Rr_a\, LC.}$$

Now a voltage variation of V_0 across the valve gives a current variation of I_0, hence—

$$r_a = \frac{V_0}{I_0} = \frac{2V_0}{i_{max}}$$

$$\therefore \quad L_1 = \sqrt{\frac{2V_0 RLC}{i_{max}}} = \frac{1}{\omega}\sqrt{\frac{2RV_0}{i_{max}}}. \quad \ldots\ldots\ldots\ldots \quad (4)$$

This formula gives the amount of inductance which should be included in one arm of the parallel circuit to give maximum efficiency. As would be expected, it is inversely proportional to the frequency of the tuned circuit.

The point A in Fig. 13, at which contact is made with the inductance and the anode lead, is known as the ANODE TAPPING POINT, and the formula above gives the value of the inductance to be included between the A.T P. and the filament lead for maximum efficiency. In other words, the formula determines the BEST POSITION OF THE ANODE TAPPING POINT UNDER SINOIDAL CONDITIONS.

The formula obtained for the correct position of the anode tapping point under sinoidal conditions is not applicable for Class " B " or Class " C " operation, but it is evident that the amplitudes of oscillatory anode current and anode voltage are still interdependent. Adjustment of the impedance of the external anode circuit without altering its frequency is still necessary, in order to obtain the most efficient power output. In other words, the necessity for an anode tapping point is unaffected, but its position at any particular frequency differs from that which would be calculated from the formula given above.

In actual practice, only rough calculations are made from formulæ of this nature, the best settings being found by trial and error.

In the case of the " divided circuit " (paragraph 8), the big condenser splits the inductance into two parts, the position of the condenser determining the anode tapping point, and the grid exciting voltage.

19 **Coupling Necessary for the Maintenance of Oscillations.**—Considering the particular case of Fig. 14, which is simply that of Fig. 3 with the addition of an anode tapping point, we have already investigated the phase relations necessary for self-oscillation, viz., that V_g and I_a are approximately in phase and V_a is approximately 180° out of phase with either. We have also found the amplitude of oscillatory current in the circuit, and the power dissipated in the circuit when the valve acts efficiently as an oscillator.

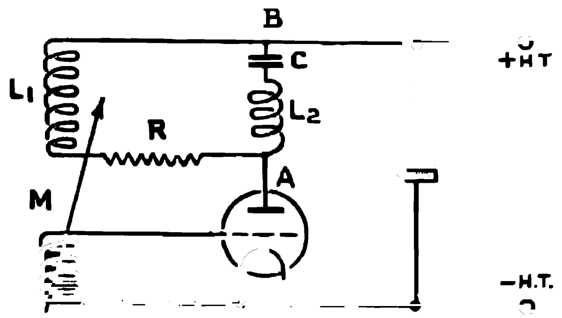


FIG. 14.

It is obvious, by analogy with previous cases, that a certain minimum coupling is necessary for self-oscillation to take place, so that, with an oscillatory action started in the circuit, the resulting anode current variation, regarded as a make-up current through the parallel circuit, is of sufficient amplitude to maintain the " circulating," or oscillatory current, at a constant amplitude.

Let us take the case of a circuit with anode tap such that the inductive branch between A and B has inductance L_1.

Let the oscillatory current (R.M.S. value) flowing in this inductance be I, the oscillatory component of anode current be I_a, and of anode voltage be V_a. Then, under oscillatory conditions, the grid-filament voltage is varied, resulting in the variation I_a above. This current variation is, however, dependent on another factor, the change in anode voltage, which it itself sets up. The general formula for I_a must therefore be used :—

$$I_a = g_m V_g + \frac{1}{r_a} V_a \quad \ldots\ldots \quad \text{(Class " A " conditions only).}$$

Now, in the circuit shown,

$$V_g = \omega MI \quad \text{and} \quad V_a = -\omega L_1 I$$

(V_g and I_a are approximately in phase, V_a in anti-phase, hence the negative sign).

$$\therefore \qquad I_a = g_m \omega MI - \frac{1}{r_a} \omega L_1 I.$$

The power given to the LC circuit to maintain oscillations

$$= I_a V_a \qquad \text{(paragraph 10)}.$$

Therefore, to maintain oscillations, the product $I_a V_a$ (numerically) must be at least equivalent to the power dissipated in the circuit, I^2R. (Neglecting damping due to grid current—paragraph 20.)

$$\text{Hence} \qquad \left(g_m \omega MI - \frac{1}{r_a} \omega L_1 I\right) \omega L_1 I \gtreqless I^2R.$$

$$\omega^2 \left(Mg_m - \frac{L_1}{r_a}\right) L_1 \gtreqless R.$$

$$\text{Now} \qquad \omega^2 = \frac{1}{LC}, \text{ where } L = L_1 + L_2$$

$$\therefore \qquad \left(Mg_m - \frac{L_1}{r_a}\right) L_1 \gtreqless LCR,$$

$$\therefore \qquad M \gtreqless \frac{CRLr_a + L_1^2}{L_1 g_m r_a} \gtreqless \frac{LCRr_a}{mL_1} + \frac{L_1}{m} \qquad \ldots\ldots\ldots\ldots\ldots\ldots \text{(1)}$$

In the case where all the inductance is on one side of the circuit, and therefore $L_1 = L$, this reduces to

$$M \gtreqless \frac{RCr_a}{m} + \frac{L}{m} \gtreqless \frac{L + CRr_a}{m} \gtreqless \frac{CR}{g_m}\left(1 + \frac{L}{CRr_a}\right) \qquad \ldots\ldots\ldots\ldots \text{(2)}$$

Now the slope of the dynamic characteristic was seen in Section "F" to be given by—

$$g'_m = \frac{m}{Z + r_a} \qquad \text{where } Z = \text{the valve external impedance} = \frac{L}{CR}$$

where m = the valve amplification factor.

$$\text{Hence} \qquad g'_m = \frac{g_m}{1 + \frac{Z}{r_a}} = \frac{g_m}{1 + \frac{L}{CRr_a}}$$

and from this, equation (2) simplifies to

$$M \gtreqless \frac{CR}{g'_m} \qquad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \text{(3)}$$

The last formula—(3)—above is exactly the same as that obtained for the maintenance of oscillations in the case of the simple oscillator used for heterodyne purposes (paragraph 4), the only difference being that in the case of the simple oscillator, the effect of the anode impedance on the anode voltage and current was neglected *i.e.*, the static characteristic was used instead of the dynamic one; in that case the formula involved g_m instead of g'_m.

If M is equal to the above expression, oscillations are just maintained; if greater, oscillations build up; if less, oscillations set up in the oscillatory circuit die away more slowly than if no reaction were employed.

Formulæ (3) and (1) may also be very simply derived by regarding the valve oscillator as an amplifier providing its own input, utilising the formula for VOLTAGE AMPLIFICATION FACTOR which is proved in Section "F."

We have V.A.F. $= m\left(\dfrac{Z}{Z + r_a}\right)$, where Z equals the external impedance, and in the case where there is no inductance in the capacity arm—

$$\text{V.A.F.} = \frac{V_a}{V_g} = m\frac{\dfrac{L}{CR}}{r_a + \dfrac{L}{CR}} = \frac{g_m r_a \cdot L}{CRr_a + L}$$

Now $$V_a \doteqdot \omega L I_L \quad \text{and} \quad V_g = \omega M I_L$$

$$\therefore \quad \frac{\omega L I_L}{\omega M I_L} = \frac{g_m r_a \cdot L}{CRr_a + L} \quad \text{or} \quad \frac{L}{M} = \frac{g_m r_a \cdot L}{CRr_a + L}.$$

$$\therefore \quad M = \frac{CRr_a + L}{g_m r_a} = \frac{CR}{g_m}\left(1 + \frac{L}{CRr_a}\right)$$

$$= \frac{CR}{g'_m} \quad \ldots\ldots\ldots \quad \text{as before.}$$

With some of the inductance in the capacity arm, formula (1) can be derived by using the relation $Z = \dfrac{L_1^2}{LCR}$.

In the case of the Hartley circuit of Fig. 5 (*b*), which is working under *sinoidal conditions*, *i.e.*, without a grid condenser and leak, the minimum value of the inductance between grid and filament for the maintenance of oscillations may be simply deduced from formula (1) above, obtained for the case of mutual inductive coupling. Calling the inductance between grid and filament L_2, it is obvious that the only difference in the argument of this paragraph is that L_2 takes the place of M, and so the formula becomes

$$L_2 = \frac{LCRr_a}{mL_1} + \frac{L_1}{m}.$$

It may be observed that the best value of L_2 for high efficiency, *i.e.*, when a grid leak and condenser are inserted, may be two or three times as great as this. The same observation applies to the case of a high efficiency Class " B " or Class " C " oscillator, in which the necessary grid excitation comes from a mutual inductive coupling. The reason is easy to find.

The coupling necessary for self-oscillations to build up until the amplitude of the grid voltage variation covers the whole of the straight part of the (I_a, V_g) dynamic characteristic, must be greater than when the operating point is half-way up the characteristic. Moreover, M, by the above theory, must equal at least CR/g'_m.

In that case, the effective value of g'_m is given by the total change in anode current divided by the total change in grid volts (Fig. 9), and this is much smaller than the slope of the dynamic characteristic, which forms the denominator of the above fraction appropriate to Class " A " conditions.

The best value of M may easily be two or three times as great as that deduced above for sinoidal conditions.

If the coupling is increased beyond this value, the amplitude of oscillatory grid voltage will increase, but the larger grid current flowing will then cause the mean potential of the grid to become more negative, so that the peak of grid voltage during the positive half cycle returns to approximately the same potential as before. Anode current, however, now flows for less than half a complete cycle, and the power input to the oscillatory circuit falls off. With increasing coupling beyond the best value, the amplitude of the oscillations thus decreases, and the grid potential may become so negative that sufficient power is not available to maintain self-oscillatory conditions, and " squegging " will occur. (*Cf.* paragraph 15).

20. **The Building Up of Oscillations ; Amplitude Reached.**—The analysis of the last paragraph is of assistance in explaining, precisely, the factors limiting the amplitude of the oscillatory anode current I_a.

Let us assume that the coupling M—or whatever provides the grid excitation—has been adjusted to some value slightly greater than the minimum required to maintain oscillations. In the general case, with the working point somewhere on the straight part of the dynamic characteristic, oscillations will build up until the UPPER OR LOWER BEND OF THE DYNAMIC CHARACTERISTIC IS REACHED, and, if the coupling is sufficient to cause further increase in amplitude, a limiting point is reached because g'_m becomes effectively less as the excursion of grid voltage past the bend increases. Finally, an amplitude is reached at which the inequality between M and CR/g'_m becomes an equality, and oscillations are maintained at a constant amplitude.

Consider the special case of a valve with steady grid and anode voltages to be such that it is working at a point on the straight part of the dynamic characteristic about half-way up ; a pictorial representation of the growth of oscillations is given in Fig. 15 It may also be assumed that the impedance of the external circuit has been adjusted so that $\mathcal{V}_a$ has its theoretical maximum value of V_0, in which case we have the conditions for both 50 per cent. efficiency, and maximum power.

In Fig. 15, the lines of maximum slope represent static characteristics.

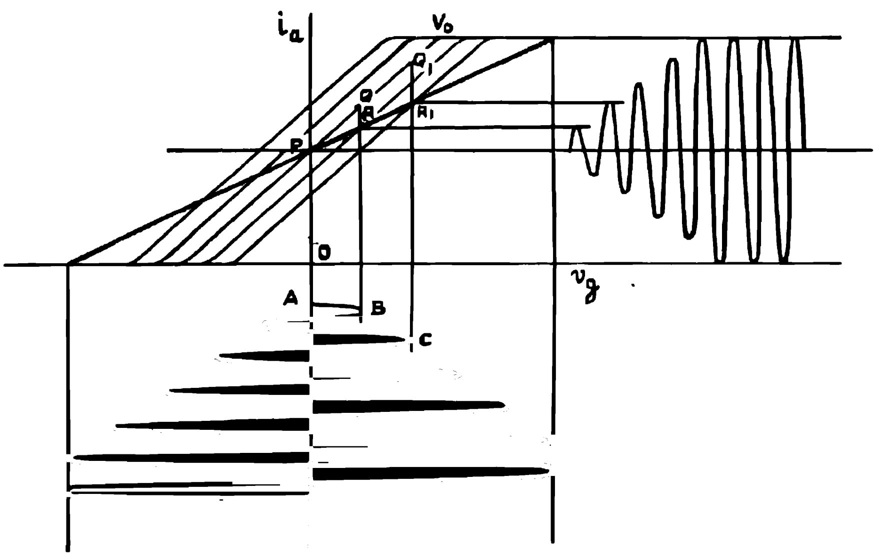


FIG. 15

An oscillation started in the LC circuit varies V_g, and consequently the anode current. Due, however, to the load in the external anode circuit, this increase in anode current is accompanied by a decrease in anode voltage, so that a positive grid swing AB does not cause the anode current to rise from P to Q, but to some value R on a different static characteristic curve drawn for the reduced anode voltage.

The next and bigger grid swing, to C, causes an increase of anode current to R_1 only, instead of to Q_1, R_1 being on a curve corresponding to a further reduced anode potential. In this way, the amplitude of anode current keeps increasing, but more slowly than if the external impedance were neglected.

Finally, the anode current builds up until it reaches the saturation value at the top or bottom bends of the DYNAMIC CHARACTERISTIC. It should be evident that P R R_1, etc., are points

on the dynamic characteristic, derived from the static by taking the impedance of the external circuit into account, and, so long as we remain on the straight part of this characteristic, conditions are exactly the same as at the start, *i.e.*, oscillations tend to build up. Once the top or bottom bend is reached, increased variations in grid voltage no longer produce correspondingly increased variations in anode current, and the amplitude of the oscillation becomes limited, mainly due to the decrease in the effective value of g'_m; M ceases to be greater than CR/g'_m.

Another factor which tends to limit the increase of amplitude of oscillations is the power consumed in the grid circuit. So far, this has been ignored, but it is obvious that with large grid voltage swings on the positive side of zero grid volts, quite considerable currents will flow in the grid circuit and cause power absorption.

21. **Classification of Valve Transmitter Circuits.**—As has already been observed (paragraph 7), there are very many different types of circuit which can be used in valve transmitters. Any attempt at classification must necessarily be of an arbitrary nature, but, in Service practice, it is found to be convenient to classify SELF OSCILLATORY CIRCUITS by means of some of their major differences.

The following is a list of the main features used in the classification :—

(*a*) The position of the tuned circuit with reference to the electrodes of the valve.
(*b*) The means by which the grid excitation is achieved.
(*c*) The method by which the H.T. voltage is applied between the anode and filament of the valve, usually called the "feed."
(*d*) The position of the electrode which is at a high oscillating potential.
(*e*) The nature of the aerial excitation.
(*f*) The arrangement of the valves in a multi-valve transmitter.

It is proposed to illustrate the use of this classification with reference to particular circuits.

22. **Position of the Tuned Circuit.**—The tuned circuit may be connected between any pair of the three electrodes in a triode :—

(*a*) The TUNED CIRCUIT may be between GRID and FILAMENT. This is generally the case in the local oscillators used in heterodyne receivers, but is not common in transmitting circuits. Fig. 1 is an example of this.
(*b*) The TUNED CIRCUIT may be between ANODE and FILAMENT. An example of this may be seen in Fig. 3.
(*c*) The TUNED CIRCUIT may be between ANODE and GRID. This type of circuit is illustrated in Figs. 5 and 6.

23. **The Grid Excitation.**—The nature of the grid exitation is another convenient feature which serves to classify transmitter circuits. There are various methods by which there is obtained an oscillatory P.D. between grid and filament in the correct phase and of sufficient amplitude for the maintenance of self-oscillations. In general, the grid exciting voltage is fed into the grid-filament circuit, by a coupling which may be DIRECT or MUTUAL, each of these headings being sub-divided into INDUCTIVE or CAPACITIVE.

Thus, the grid excitation in the case of Fig. 3 is obtained by MUTUAL INDUCTIVE COUPLING. In the Colpitts circuit (Fig. 7), the grid excitation would be described as DIRECT CAPACITIVE, the tuned circuit being between anode and grid.

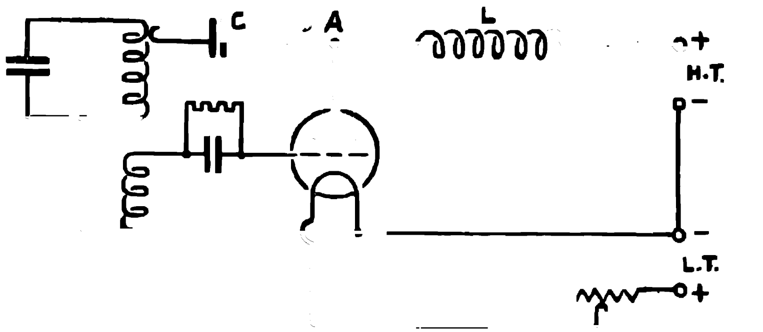
In the Hartley circuit (Fig. 5), the grid excitation is DIRECT INDUCTIVE.

24. **The Nature of the Feed.**—This should not be confused with the methods whereby the large H.T. voltages required for transmitting valves are produced.

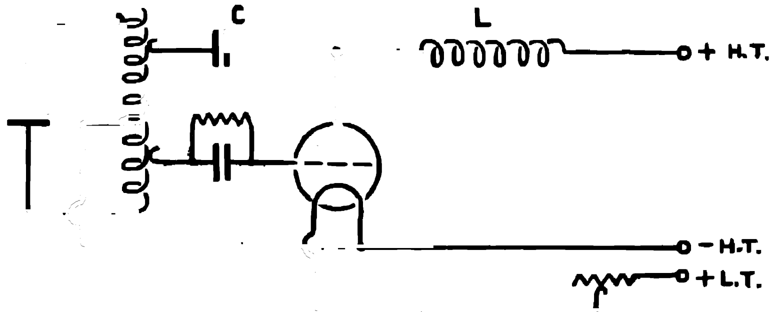
Two types of feed are commonly used in valve transmitting circuits :—

(*a*) SERIES FEED.—The H.T. supply, the valve and the tuned circuit are in series. Examples of this have been seen in Figs. 3, 5 and 6.

(*b*) PARALLEL FEED.—The valve and the tuned circuit are in parallel across the H.T. supply. Examples of this are shown in Fig. 16.



(a)



(b)

FIG. 16.

In the case of PARALLEL FEED, two additional components are rendered necessary, the choke L, known as the ANODE CHOKE, and the condenser C, known as the ANODE BLOCKING CONDENSER.

The anode blocking condenser is necessary to prevent a short circuit of the H.T. supply through the tuned circuit inductance. It is large enough to be regarded as of negligible impedance to radio frequency currents.

The function of the anode choke is as follows :—

The H.T. supply is in parallel with the tuned circuit across the valve. Being of comparatively low resistance, it would constitute practically a short circuit for the tuned circuit as regards oscillatory currents, and introduce such damping losses as would effectually prohibit the maintenance of oscillations. The introduction of a large choke in series with the H T. supply increases the impedance, to radio frequency currents, of this parallel path sufficiently to render its damping effect

negligible, and operates so as practically to confine the radio frequency currents to the valve and tuned circuit. There will, of course, be a small radio frequency current through the choke and H.T. supply, corresponding to the oscillatory P.D. across the tuned circuit and valve. In practice, the H.T. supply will be paralleled by a large condenser to by-pass this small radio frequency current.

The choke thus enables the anode to adjust itself to its necessary radio frequency voltage changes, without robbing the tuned circuit of excessive radio frequency current.

There is also a mean anode current flowing through the valve, as has been shown by the analysis given in earlier paragraphs. To this the choke presents a negligible impedance.

In the case of SERIES FEED it is still more important that the H.T. supply should be by-passed by a large condenser, for otherwise the whole of the oscillatory make-up current would flow through it.

Two methods of inserting the feed, and its by-pass condenser, have already been seen in Figs. 5 and 6.

25. **The High Oscillating Potential Electrode.**—In Fig. 17, the H.T. positive terminal is directly connected to the anode. As in Fig. 6 (*a*), the condenser C is a large by-pass condenser whose reactance at radio frequencies is negligible. In consequence, the point A and the anode may be regarded as common, when oscillatory potentials only are considered. The D.C. potentials of the anode and the point A are, of course, very different; the steady potential of the anode is that of the H.T. supply, and, since the point A is earthed, there can be practically no oscillatory variation of anode potential. The electrode whose potential actually undergoes the large oscillatory variation (of amplitude approaching the H.T. voltage when the transmitter is adjusted efficiently) is the filament. This type of feed accordingly leads to another important practical distinction between transmitters.

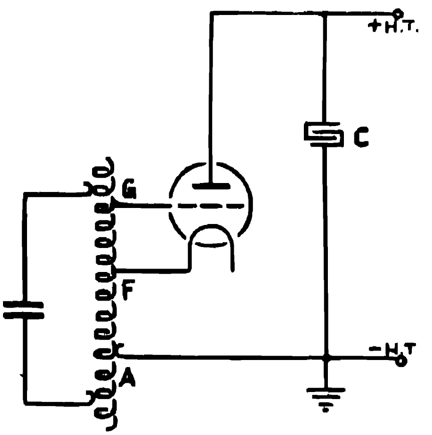


FIG. 17.

(1) Transmitters in which the ANODE POTENTIAL UNDERGOES OSCILLATORY VARIATION OF LARGE AMPLITUDE. This has been the case in all the previous types discussed.

(2) Transmitters in which the LARGE OSCILLATORY POTENTIAL VARIATION OCCURS AT THE FILAMENT, as in the case just considered. This involves, in practice, that the filament heating circuit must be well insulated from earth, a condition which is most easily fulfilled by an alternating heating current provided via the secondary of a transformer.

In both cases, of course, the high D.C. potential of the anode involves efficient insulation of that electrode.

The instantaneous potentials of the three electrodes in case (1) have already been illustrated in Figs. 8 and 9. These instantaneous potentials in case (2) are shown in Fig. 18 (*a*). The corresponding anode-filament and grid-filament P.Ds. for this case are exhibited in Fig. 18 (*b*) to impress the point that there is no difference in fundamental principle in the two cases. The derivation of Fig. 18 (*b*) from Fig. 18 (*a*) is obvious.

26. **Aerial Excitation.**—The consideration of the tuned circuit has hitherto been confined to the question of how an undamped radio frequency current may be maintained in it. We now pass to the question of how the oscillatory energy, thus obtained, is to be made available for radiation as electromagnetic waves into space.

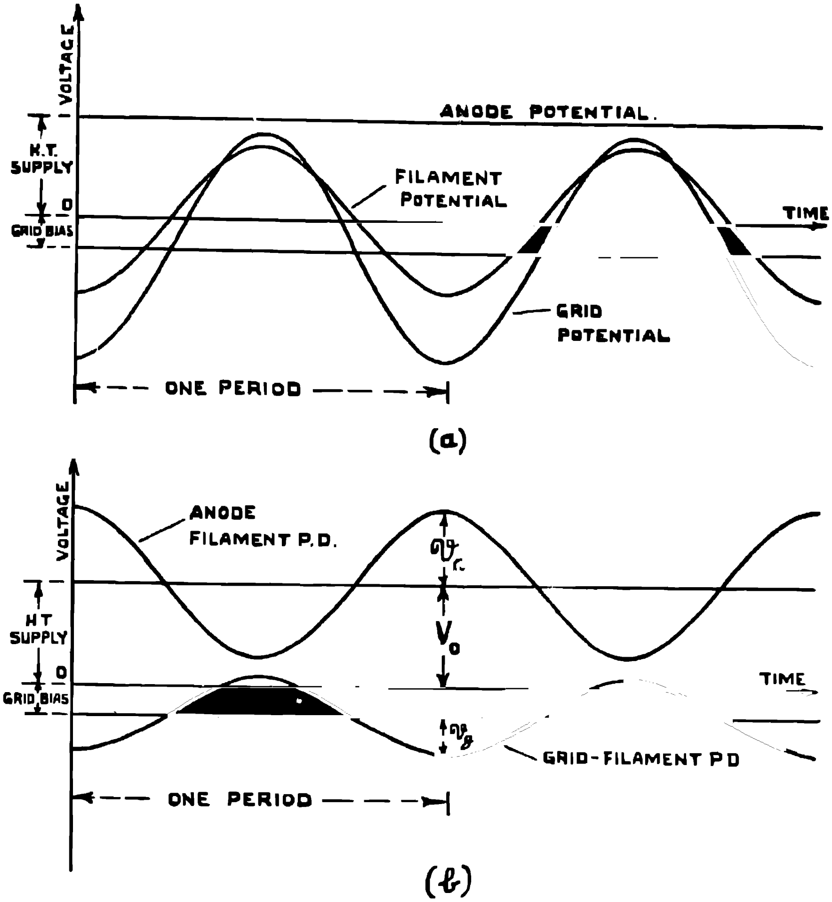


FIG. 18.

(1) The tuned circuit itself may be employed for this purpose, *i.e.*, it may be in the form of an open oscillatory circuit. The condenser then corresponds to the distributed aerial capacity to earth, and the inductance to the natural aerial inductance and artificial inductances in series. This may be called **direct** aerial excitation.

A Service transmitting circuit of this type is shown in Fig. 19. With reference to the points of transmitter design discussed above, its description is as follows :—

Tuned circuit between anode and grid.
Direct inductive grid excitation.
Series feed.
Filament at high oscillatory potential.
Direct inductive aerial excitation.

In this transmitter, efficiency has, to some extent, been sacrificed for simplicity of adjustments, as no provision is made for varying the grid excitation independently.

The 0·25 jar condenser in series with the aerial capacity is the ordinary series condenser for higher frequencies. The filament heating circuit has been omitted to simplify the diagram. As previously remarked, the heating circuit must be adequately insulated.

(2) The aerial circuit may be a separate open oscillatory circuit, coupled to the valve tuned circuit, which is then a closed oscillator. This may be described as **mutual** aerial excitation. The coupling is usually mutual inductive.

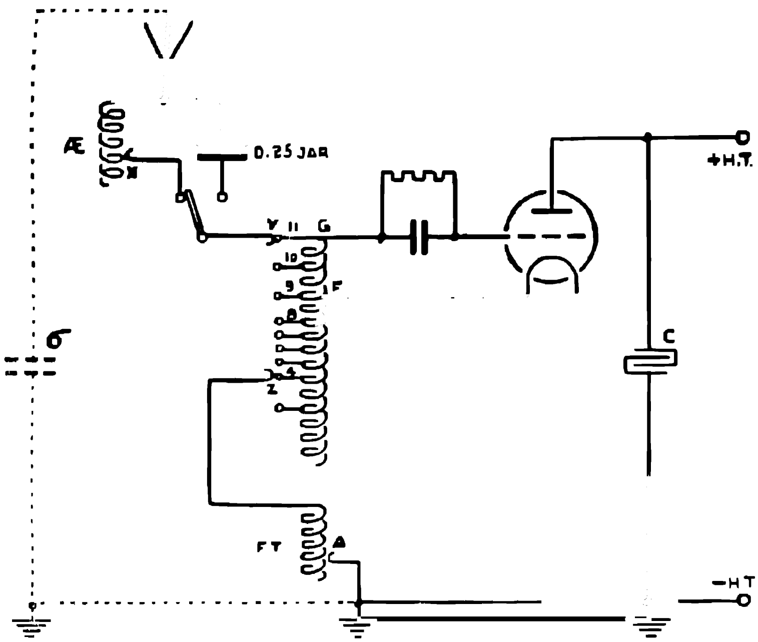


FIG. 19.

Fig. 20 shows a transmitting circuit of this nature.

This circuit should be classified as follows :—

Tuned circuit between anode and filament.
Mutual inductive grid excitation.
Parallel feed.
Anode at high oscillatory potential.
Mutual inductive aerial excitation.

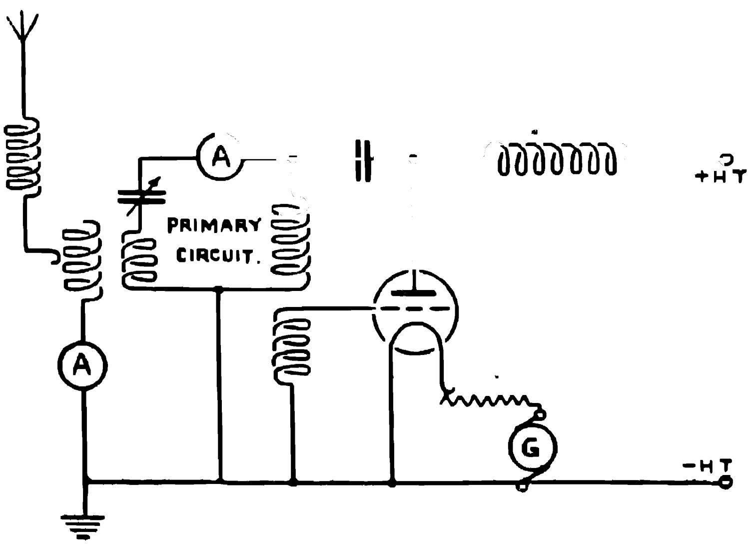


FIG. 20.

27. **Mutual Aerial Excitation.**—The choice between a closed oscillatory circuit with coupled aerial, and an open oscillatory circuit directly excited by the valve depends on certain factors, now to be examined. Mutual aerial excitation has the following **advantages** :—

(*a*) Valves are constructed so as to be capable of dissipating a certain amount of power at the anode with safety. This is known as the **valve rating** (*see* Section "B"). The insulation of the aerial (if an open oscillatory circuit is employed) may sometimes fall to a very low value. This may occur in submarines with the deck insulator "washing over," or in surface ships during heavy rain, and the load thus thrown on the valve may be so great that the aerial stops oscillating. Under these non-oscillatory conditions, the whole of the power taken from the H.T. supply is dissipated at the anode of the valve, instead of a percentage of it (say 30 per cent. to 50 per cent.), and the valve may be burnt out. If, however, a mutually coupled aerial circuit is used, the closed circuit continues to oscillate and to absorb its share of the power. The closed circuit is always trying to produce forced oscillations in the aerial, and hence tends to dry off the moisture from the aerial insulators.

(*b*) The inevitable occurrence of harmonics in a self oscillatory circuit with negative grid bias has already been emphasised. With direct aerial excitation, energy at the frequency of these harmonics is radiated without any limitations, and produces undesirable interference.

When mutual excitation is employed, the aerial circuit is naturally tuned to the first harmonic, and so presents considerably less impedance to the E.M.F. induced in it at this frequency than it does to the E.M.F.s induced by the higher harmonic currents flowing in the closed oscillatory circuit. Thus, the higher harmonic currents in the aerial circuit are considerably reduced in amplitude compared with the first harmonic current, and correspondingly less energy is radiated at these unwanted frequencies.

In addition, the proportion of harmonic to fundamental current flowing in the inductive branch of the primary circuit is considerably less than the corresponding proportion in the make-up (anode) current, for the capacitive branch of the primary tuned circuit presents less reactance to these higher frequency currents than does the inductive branch.

(*c*) The question of the variation of the natural frequency of self-excited oscillations will be considered in greater detail when master oscillators are discussed. It may be pointed out now, however, that one common cause of such variation arises through changes in aerial capacity. These changes may be due to the aerial swaying with the wind, or to alterations in the "earth," such as are occasioned by training a turret on board ship, or proximity in dock to moving machinery, *e.g.*, cranes and sheer legs. With direct aerial excitation these changes in capacity are directly operative in altering the LC value of the oscillatory circuit, and therefore the transmitted frequency. They also alter the $\frac{L}{C}$ ratio of the tuned circuit, and so affect the oscillatory power transferred from the valve.

With a closed oscillatory circuit and coupled aerial, variations in aerial capacity still operate indirectly in the same way, but their effect then depends also on the coupling factor, as may be seen from the discussion in Vol. I of oscillations in coupled circuits. With loose coupling, the variations in frequency due to this cause may therefore be greatly reduced. Mutual aerial excitation is particularly advantageous at high frequencies where an untuned coupled aerial is usually employed. In this case, variations in aerial capacity have a completely negligible effect With direct aerial excitation, variations in aerial capacity would then produce the greatest actual changes in the transmitted frequency, the percentage variation remaining the same.

The obvious **disadvantage** of mutual aerial excitation is that two circuits are maintained in oscillation with a corresponding increase in damping losses. Also, the looser the coupling, the smaller is the amount of energy transferred to the aerial circuit. Nevertheless, fairly loose coupling must be employed, one reason being the advantage in minimising frequency changes as mentioned above. Of more importance, however, is the avoidance of a phenomenon known as "**frequency jump,**" which occurs in valve transmitters with tightly-coupled aerial circuits.

This may be compared with the double frequency effect in coupled spark transmitters. The distinction is that a coupled valve transmitter cannot oscillate on both these frequencies at once, though it may oscillate on either, according to the total equivalent impedance of the two tuned circuits. The valve oscillates on whichever of the two frequencies corresponds to the lower equivalent impedance.

It is impossible, however, to keep the primary and aerial circuits exactly in tune. The natural frequency of the primary circuit depends, to some extent, on the valve constants and the potentials of the electrodes, and that of the aerial circuit varies with the aerial capacity. The difficulty of keeping all these factors constant produces the result that both the primary and aerial circuits vary to some extent about the frequency to which they are both originally tuned, and so the equivalent impedance also varies. Thus, when oscillations are taking place on one of the two possible frequencies, the variation of equivalent impedance may result in the impedance at the other possible frequency becoming less than that of the actual frequency of oscillation, although it was previously greater. The frequency of oscillation then suddenly alters to this other possible value, giving a "frequency jump" equal to the difference between the two possible frequencies. This is accompanied by sudden changes in the values of the primary and aerial currents, which enable the effect to be detected.

This phenomenon is only noticeable with tight coupling, where there is a large difference between the two possible frequencies of oscillation. It is obviously undesirable, and should be avoided by always employing loose coupling.

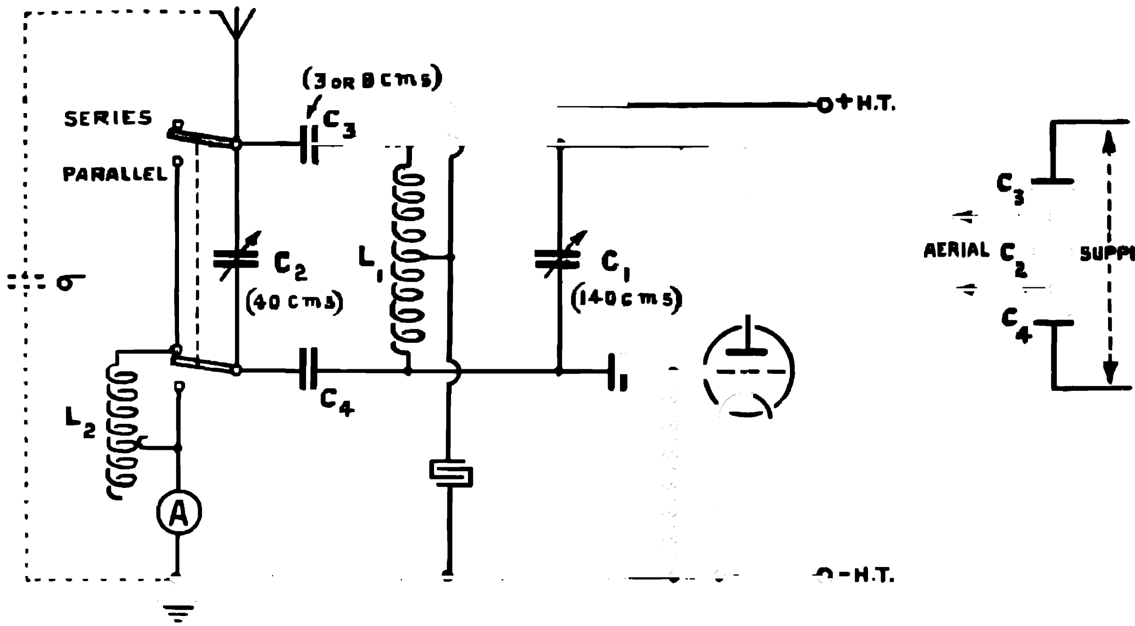
28. **Indirect Capacitive Aerial Excitation.**—For certain purposes, capacitive aerial excitation has advantages. In practice, with a variable mutual inductive coupling, it is not easy to avoid accidental over-coupling, with consequent risk of "frequency jump." It is quite simple to design a capacitive coupling with the requisite small value of coupling coefficient, and with the necessary rigidity in construction to provide foolproof operation, to obviate any chance of accidental over-coupling.

Fig. 21 (*a*) gives the circuit details of a typical H/F transmitter employing "indirect capacitive aerial excitation," condensers C_3 and C_4 being the condensers which are not common to either circuit. It will be seen that the tuned circuit is of the series fed Hartley type, the tuned circuit being between anode and grid.

In some cases the coupling condenser C_4 is omitted.

The circuit also shows a switch by which coil L_2 may be put either in series or in parallel with the variable condenser C_2. The necessity for this arrangement arises from the fact that at H/F, when the aerial may often be equivalent to an even number of quarter waves in length, it is frequently necessary to feed the aerial across the high impedance coupling which is provided by a rejector circuit (Section "R" 20). The coupling condensers C_3 and C_4 have very small capacities, and, in one Service example, have values of 3 cms. at the upper end, and 8 cms. at the lower end of the H/F range.

A certain technique is required in tuning a transmitter, coupled to an aerial in this way. The main oscillatory circuit should be tuned by using—where there is a choice—a relatively small value of L_1, and a relatively large value of C_1; C_1 is in parallel with the anode-grid capacity, and, by making the former large, one tends to eliminate frequency variation, by "swamping" any fluctuations in the inter-electrode capacity. The aerial circuit is tuned by trying combinations of L_2 and C_2, together with the series-parallel switch, until maximum current reading is obtained in the



(a) (b)

FIG. 21.

ammeter A. The size of condenser C_2 determines the amount of energy transferred from the primary to the aerial circuit. For a given frequency, therefore, the smallest value of aerial capacity, and the largest value of aerial tuning inductance L_2, should be used, in order to obtain optimum results. If C_2 is made too big, its reactance becomes relatively small; the oscillatory voltage across it will be much reduced and practically nothing will be fed into the aerial. Fig. 21 (*b*) demonstrates that the aerial is tapped off C_2, and shows that if its impedance is made too small, the aerial is, virtually, "shorted."

Although the value of C_2 must be relatively small, it is evident, however, that C_2 cannot be decreased beyond a certain value if over-coupling is to be avoided and the risk of frequency jump eliminated. The coupling factor of the circuit is given by—

$$K = \frac{C'}{\sqrt{(C_2 + C')(C_1 + C')}} \qquad \text{where} \qquad C' = \frac{C_3 C_4}{C_3 + C_4}.$$

In practice, it is necessary to have some " test " for determining the absence of frequency jump. The presence of this phenomenon is usually indicated by sudden alterations of the aerial current given by ammeter A. A slow alteration of the value of condenser C_2, or C_1, should produce a simultaneous slow alteration of aerial current. If this is not the case, frequency jump should be suspected, and another setting of the combination of L_2 and C_2 should be tried.

★29. OVER-COUPLING ; ZIEHEN OR FREQUENCY JUMP EFFECT.—Frequency jump is an interesting phenomenon and it is worth while investigating the matter a little more closely. When a self-oscillatory circuit is coupled to another tuned circuit, Fig. 22 (*a*), the frequency generated corresponds almost exactly to the resonant frequency of the combination, or frequency at which the whole circuit has zero reactance. Unfortunately, there are two of these frequencies ; this arises from the same cause that accounts for the double frequency effect seen in spark transmitters, or the double humped current/frequency curve, obtained when an E.M.F. of steady R.M.S. value is applied to one circuit coupled to another.

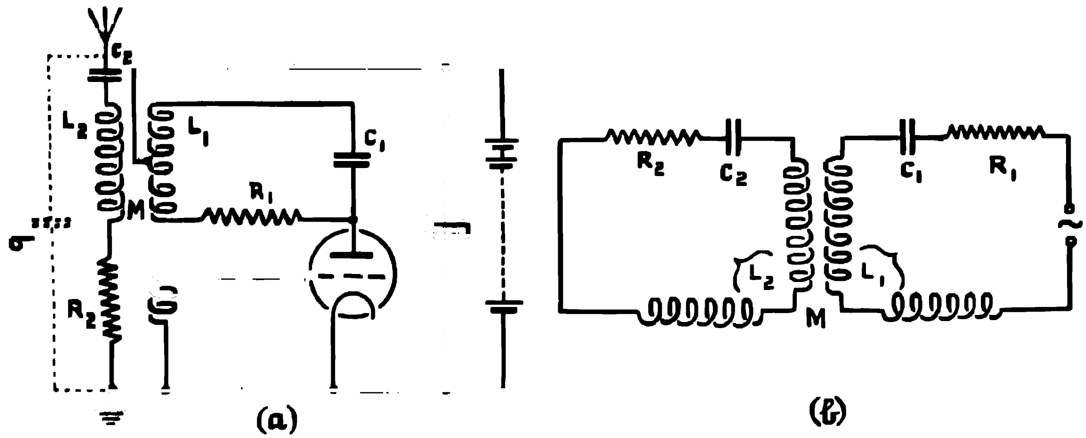


FIG. 22.

Now in Volume I, the case of Fig. 22 (*b*) was analysed ; the reactance term in the nett impedance expression was equated to zero, and by neglecting R_2 in comparison with X_2, it was shown that

$$X_1X_2 = \omega^2M^2. \qquad \text{(1)}$$

If the two circuits are each tuned to the same frequency $\frac{\omega_0}{2\pi}$, this leads to the well-known formula $\omega = \frac{\omega_0}{\sqrt{1 \pm K}}$, proved in Volume I.

In the more generally experienced case, the two circuits are not each in tune, hence—

From (1) .. $\left(\omega L_1 - \frac{1}{\omega C_1}\right)\left(\omega L_2 - \frac{1}{\omega C_2}\right) = \omega^2M^2.$

Now $\omega_1^2 = 1/L_1C_1$ and $\omega_2^2 = 1/L_2C_2$.

$$\therefore \quad \left(1 - \frac{\omega_1^2}{\omega^2}\right)\left(1 - \frac{\omega_2^2}{\omega^2}\right) = \frac{M^2}{L_1L_2} = K^2, \qquad \text{(2)}$$

where K is the coupling factor. Equation (2) may also be expressed in the form—

$$(1 - K^2)\,\omega^4 - (\omega_2^2 + \omega_1^2)\,\omega^2 + \omega_1^2\,\omega_2^2 = 0. \qquad \text{(3)}$$

In the above equation, it will be supposed that ω_1^2 is maintained constant and that ω_2^2 is the " independent variable " which alters as (say) L_2 is altered. Equation (3) then becomes A RELATION BETWEEN ω^2 AND ω_2^2. The graphical interpretation is a hyperbola, Fig. 23 (*a*). For any value of ω_2^2, there are two possible values of ω^2, the resonant frequency of the whole system.

In general, the current will tend to oscillate at that frequency which is nearest to the natural frequency of the valve oscillator, and corresponds to ω_1. At that frequency, the total equivalent impedance will be the least.

In more detail the matter may be examined as follows :—

From (3) .... when $\omega_2^2 = 0$

$$(1 - K^2)\,\omega^4 - \omega_1^2\,\omega^2 = 0$$

or $\quad \omega^2\left\{(1 - K^2)\,\omega^2 - \omega_1^2\right\} = 0,$

i.e., $\quad \omega^2 = 0$ and $\omega^2 = \frac{\omega_1^2}{1 - K^2}$

The first of these results shows that one branch of the hyperbola passes through the origin 0 ; the second value of ω^2 determines N, the point where the other branch of the curve cuts the ω^2 axis. It will be observed that when K is small, $\omega^2 \doteqdot \omega_1^2$, the resonant frequency of the whole system being practically determined by that of the oscillator circuit.

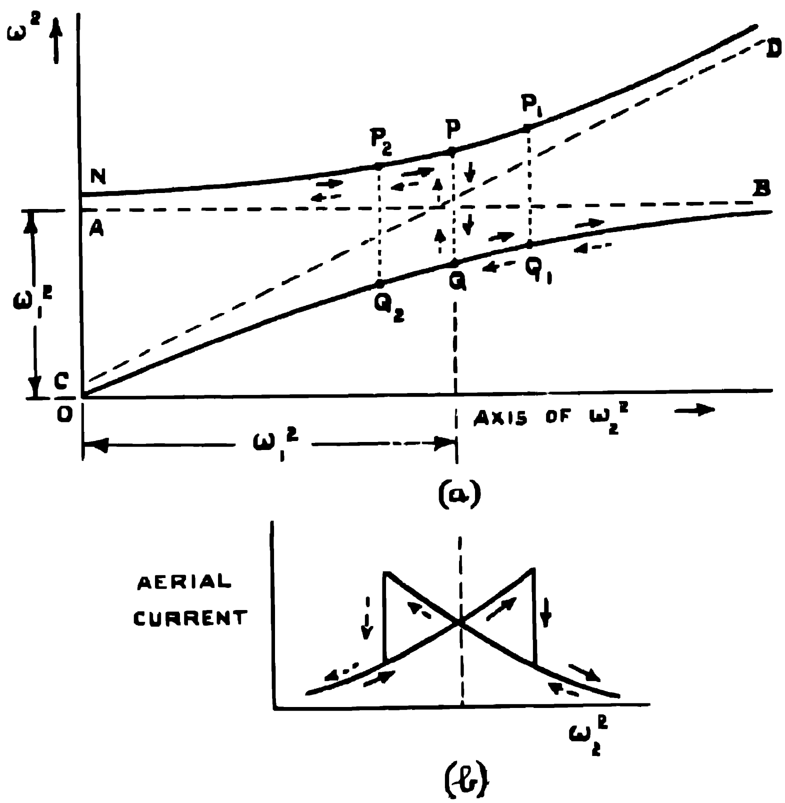


FIG. 23.

From (2) .... when $\omega_2^2 = \omega_1^2 = \omega_0^2$

$$1 - \frac{\omega_1^2}{\omega^2} = \pm K \quad \text{or} \quad = \omega^2 - \frac{\omega_0^2}{1 \pm K}$$

In this case, the two possible values of ω^2 are given by $\frac{\omega_0^2}{1+K}$ and $\frac{\omega_0^2}{1-K}$, or approximately, when K is small, $\omega_0^2(1 - K)$ and $\omega_0^2(1 + K)$. These are the points Q and P, respectively, on the graph Fig. 23 (*a*); it is clear that they are equally spaced below and above the line $\omega = \omega_0 = \omega_1$. If the value of ω^2 jumps from the one to the other, for any reason, we have the approximate relation—

$$PQ = K\omega_0^2. \qquad (4)$$

Also from (2) we see that when ω^2 approaches the value ω_1^2 FROM BELOW IT, we have $\left(1 - \frac{\omega_2^2}{\omega^2}\right) \rightarrow -\infty$, *i.e.*, $\omega_2^2 \rightarrow +\infty$; the curve asymptotes the line $\omega^2 = \omega_1^2$, for the value $\omega_2^2 = +\infty$.

When ω^2 approaches the value ω_1^2 FROM ABOVE IT, we have $\left(1 - \frac{\omega_2^2}{\omega^2}\right) \rightarrow +\infty$, *i.e.*, $\omega_2 \rightarrow -\infty$.

More generally, it may be shown that the asymptotes to the curve are given by the equations—

$$\omega^2 = \omega_1^2 \quad \text{and} \quad \omega^2(1 - K^2) = \omega_1^2 + \omega_2^2 K^2$$

The first asymptote is horizontal, and the second cuts the $\bar{\omega}^2$ axis at C where $\omega_2^2 = 0$, *i.e.*, at $\omega^2 = \omega_1^2 K^2/(1 - K^2)$.

With regard to the two possible values of ω^2, it should now be clear that they will always be above or below the line AB, *i.e.*, one will be greater and the other less than ω_1^2.

The branch of the curve which the valve oscillator will actually follow, or partially follow, depends upon the operating conditions ; it depends very largely upon whether the value of ω_2^2 is being increased from small values upwards (tuning up), or whether the value of ω_2^2 is being gradually decreased from large values downwards (tuning down).

When L_2 is big, the frequency and the corresponding value of ω_2^2 is low ; the aerial current I_2 will be very small and will produce an inappreciable effect on the value of ω_1^2. Oscillations will commence at a value of ω^2 given by the upper branch of the curve at a point on it where the frequency differs very little from the natural frequency of the oscillator, *i.e.*, corresponding very closely to the line AB which is ω_1^2.

As L_2 is decreased, the frequency is increased and the value of ω^2 follows the curve from N towards P. At the same time the aerial current increases, producing a steadily increasing effect on the primary circuit ; its effect is the same as an increase in the primary resistance, and the primary current falls to a minimum value. The impedance of the whole circuit has increased.

The points P and Q on the curves, correspond to values of ω^2 which are approximately equidistant from the line AB. At P, the impedance of the whole circuit will thus be the same for the two possible values of ω_2^2. The point will therefore be one of great frequency instability.

If the primary circuit, *i.e.*, the oscillator, does not possess much surplus " negative resistance " an increase in the value of ω_2^2 corresponding to the point P results in the lower curve coming nearer to AB than the upper one. The impedance of the whole, corresponding to values of ω^2 given by the lower curve, will then be less than that given by the upper curve, and the frequency suddenly falls, producing an alteration in ω^2 represented by PQ. Values of ω^2 corresponding to a further increase in ω_2^2 will then be given by the lower curve, which rapidly approaches the line AB.

In Fig. 23 (*a*), the full arrows trace the path NPQB, and represent the changes in ω^2 when the secondary circuit is tuned UP. The dotted arrows represent the reverse path which would be followed if the changes in ω_2^2 were made in the reverse direction, *i.e.*, tuning DOWN.

In the above process it has been assumed that the oscillator had not much surplus of negative resistance, *i.e.*, the coupling between grid and anode in Fig. 22 (*a*) is just sufficient to maintain oscillations. In the case where this coupling is in excess of that minimum value, the oscillator will possess a considerable excess of negative resistance. The effect of this will be that for increasing values of ω_2^2, the value of ω^2 will follow the upper curve beyond the point P, proceeding towards P_1. On account of the negative resistance in the circuit, frequencies represented by values of ω^2 given by the lower curve no longer give a lower impedance than that given by frequencies corresponding to the upper curve. The frequency jump is delayed until the aerial current is large enough to make the impedance of the whole circuit to the larger value of ω^2 greater than the total impedance to the lower values. If this takes place at some point P_1, the frequency will suddenly jump by an amount represented by P_1Q_1 ; from that point values of ω^2 will follow the lower curve.

Under similar conditions, with an excess of negative resistance in the primary circuit, the path followed for decreasing values of ω_2^2 would be Q_1 Q_2 P_2 N.

A sensitive indication of the presence of frequency jump is given by the alteration in reading of the aerial ammeter. Fig. 23 (*b*) represents diagrammatically the changes in aerial current with alteration of ω_2^2.

It has been seen that the aerial current is an important factor in determining which of the curves is followed ; when changes take place in the latter as, for example, during keying, trouble due to frequency jump is sometimes experienced.

30. **Practical Aspects of Ziehen Effect.**—In regard to the value of the coupling factor K, it appears, from the foregoing treatment, that there are conflicting requirements. To effect the maximum transfer of energy, the value of K should be high ; to avoid the phenomenon of frequency jump, the value of K must be maintained reasonably small. For practical purposes a compromise has to be found.

At the transmitter, the practical test for the presence of frequency jump, described in paragraph 28, is simple and easy to apply. If this phenomenon passes undetected at the transmitting end, its presence is revealed at the receiving end by the loss of part of the signal. The signal may arrive deficient in some of the " shorts " which, together with the " longs," go to make a morse code signal. The missing " shorts " will actually have been radiated, but at a different frequency, and, if the frequency jump is considerable, will not be received in the telephones of the operator at the receiving end. The signalled message is therefore unintelligible.

The remedy to be applied at the transmitting end is EITHER to decrease the coupling, OR to de-tune the aerial circuit slightly away from the points P or Q. With reference to Fig. 21 (*a*), this adjustment is effected in practice by tuning the aerial circuit to give maximum current, and then slightly turning back on the adjustment, *i.e.*, de-tuning

31. **Drying-out Circuit.**—The disadvantage arising in directly-excited aerial circuits due to intermittent wetting of the aerial insulation, and the consequent risk of cessation of oscillations and damage to the valves, may be obviated by a "drying-out" circuit such as that shown in Fig. 24.

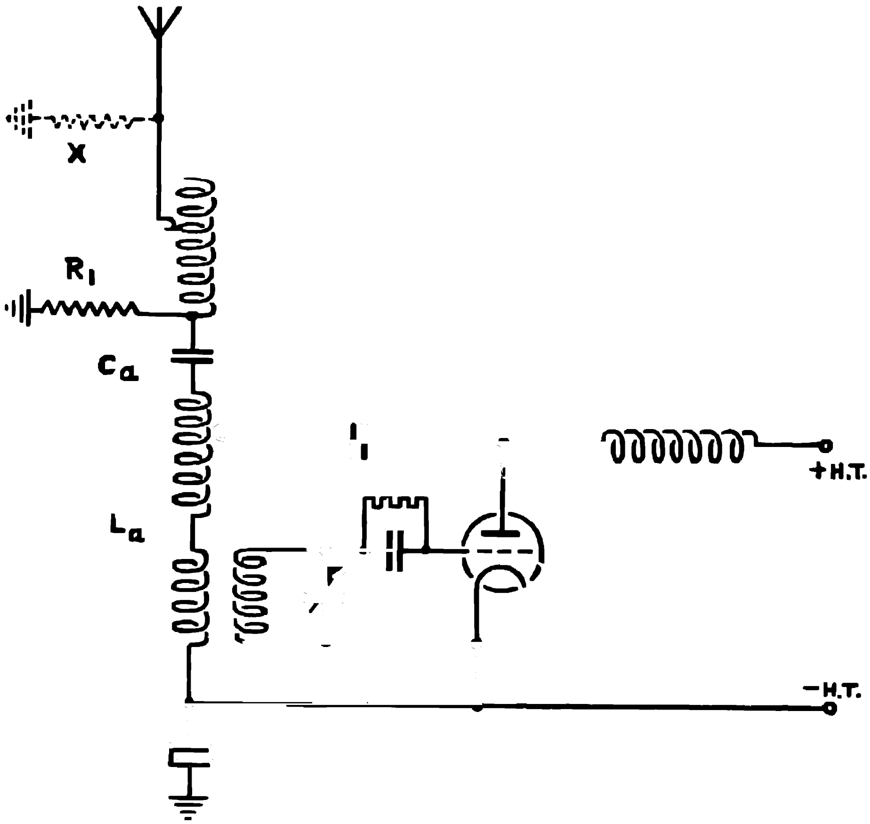


FIG. 24.

It consists of a series acceptor circuit (L_aC_a) introduced into the aerial circuit, and a connection to earth through a resistance (R_1), from the junction of these two circuits.

The introduction of the acceptor circuit L_aC_a has provided a circuit which will oscillate even if the aerial circuit is earthed at the deck insulator.

Now the aerial circuit is resistance-coupled to this circuit by reason of the P.D. across the resistance R_1.

If the deck insulator is earthed by salt water over its surface, as indicated at X, the high frequency P.D. across it will cause a leakage current to flow through this moisture, tending to dry off the insulator. At the same time, provided the aerial and acceptor circuits are in resonance, oscillations will be forced on the aerial circuit.

As the deck insulator is a point of high potential in the aerial circuit, the moisture is rapidly dried off, once the aerial oscillations are set up.

This loss by leakage is equivalent to a damping loss R_2 introduced into the aerial.

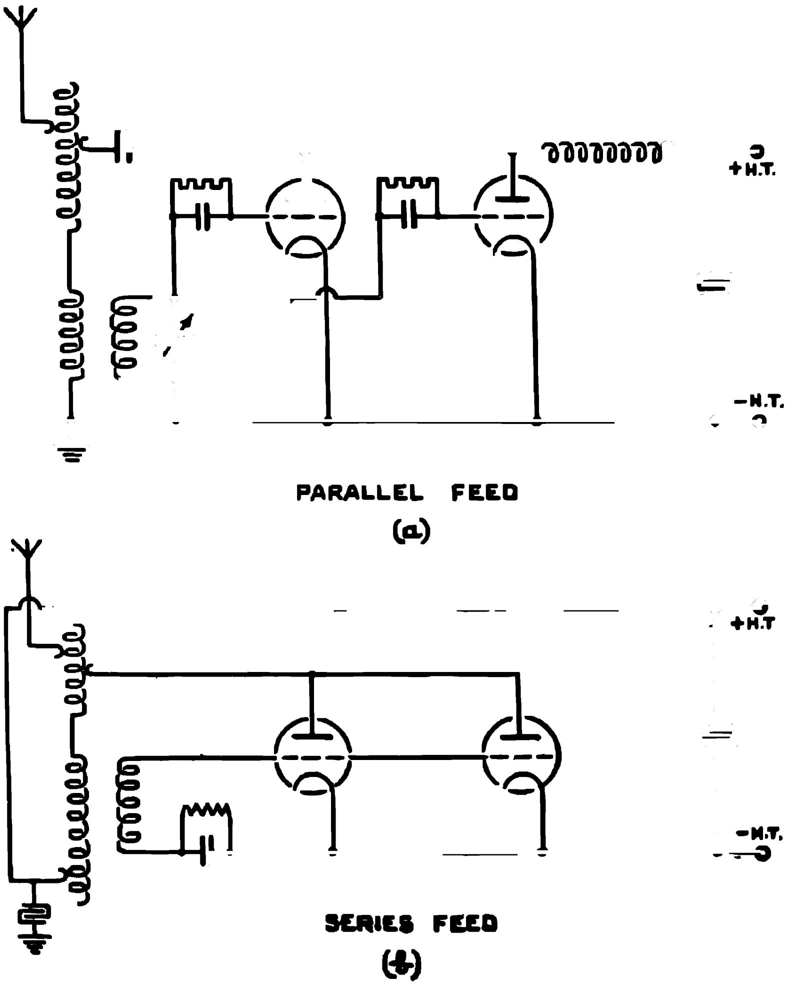
The joint resistance of R_1 and R_2 is equal to $\frac{R_1 R_2}{R_1 + R_2}$, and can never be greater than R_1, however great the aerial damping.

32. **Increasing the Power; Multivalve Arrangements.**—The heat capable of being dissipated at the anode of one valve is limited and, therefore, so is the amount of oscillatory power available for radiation. To increase the power in the aerial circuit, the number of valves must increased, and this raises the question of their relative arrangement. There are three possib valve dispositions :—

(1) In series.
(2) In parallel.
(3) In push-pull.

The first of these involves the necessity of providing an H.T. voltage supply greater than that necessary for one valve. Series circuits are little used, and the series-modulation circuit provides the best example (N.22). Parallel and push-pull arrangements in valves are, however, frequently encountered.

33. **Valves in Parallel.**—An arrangement of **valves in parallel** is illustrated in Fig. 25. The anodes are joined in parallel, with the same H.T. voltage applied to each. In Fig. 25 (*a*), parallel feed is shown, and series feed in Fig. 25 (*b*).



Transmitting Valves in Parallel.
FIG. 25.

The grids are connected in parallel to the grid excitation coil. In one case they are shown with separate condensers and leaks, in the other case this fitting is common. This depends on the nature of the valves used. The filaments are normally provided with separate rheostats, but this precaution is rapidly being dispensed with, owing to the improved similarity of performance of modern transmitting valves.

If two valves in parallel have the same valve constants r_a, g_m and m, it follows that :—

(*a*) Their joint A.C. resistance is $r_a/2$.

(*b*) Their joint mutual conductance is $2g_m$, for the same applied grid voltage gives twice the total ANODE CURRENT through the output circuit that would be obtained from either valve singly.

(*c*) The amplification factor of the combination

$$= r_a/2 \times 2g_m = r_a g_m = m,$$

and so is the same as the amplification factors of the individual valves.

(*d*) The amount of heat capable of being dissipated at the anodes is doubled, *i.e.*, the anode rating of the combination is twice that of the individual valves, ; thus the power that can be developed safely in the oscillatory circuit is doubled.

If the damping losses in the oscillatory circuit are equivalent to a resistance R, and the R.M.S. CIRCULATING CURRENT is I_1 with one valve, the oscillatory power transferred to the aerial when continuous oscillations are maintained is $I_1^2 R$.

Let the corresponding oscillatory power available with two valves in parallel be $I_2^2 R$, then since the power is doubled—

$$I_2^2 R = 2I_1^2 R \qquad \therefore \qquad I_2^2 = 2I_1^2 \text{ and } I_2 = \sqrt{2}\, I_1.$$

Thus the amplitude of oscillatory circulating current is increased $\sqrt{2}$ times.

With n similar valves in parallel, the amplitude of the circulating current is $\sqrt{n}$ times that obtained with one valve ; in this case, the "circulating current" is also the "aerial current." The decrease in A.C. resistance of this arrangement necessitates a corresponding decrease in the impedance of the oscillatory circuit to maintain the most efficient conditions, *i.e.*, the position of the anode tapping point must be altered—in general, it is lowered. Under sinoidal conditions, the formula for the amount of inductance "L_1" between anode and filament, viz., $L_1 = \sqrt{LCRr_a}$, allows the change in L_1 to be deduced.

With two valves in parallel, r_a becomes $\frac{r_a}{2}$, and therefore "L_1" becomes $\frac{L_1}{\sqrt{2}}$, *i.e.*, THE ANODE TAPPING POINT MUST BE LOWERED.

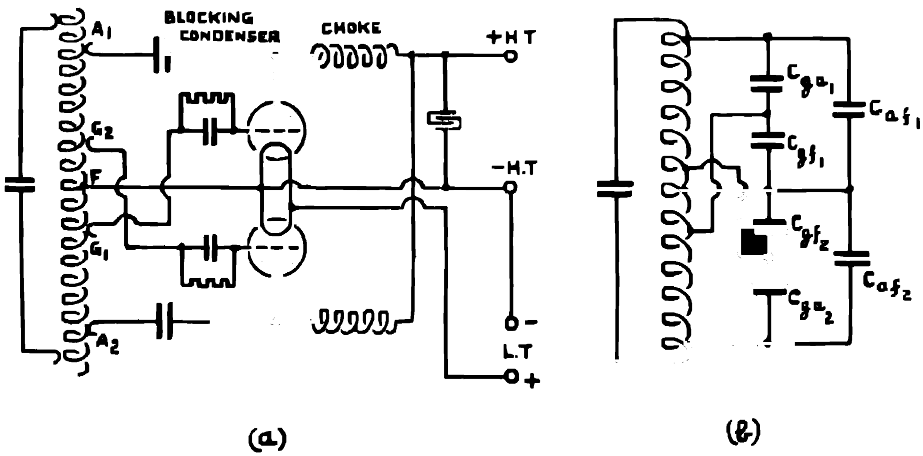
With n valves in parallel, L_1 would have to be reduced to $\frac{L_1}{\sqrt{n}}$ to preserve maximum efficiency.

Alternatively, it may be considered that as the circulating current has increased $\sqrt{n}$ times, the inductance between anode and filament must be decreased $\sqrt{n}$ times to preserve the same anode-filament P.D. $(\omega L \mathscr{I}_L)$.

It is obvious that the anode tapping point cannot be lowered indefinitely, and in practice a point is eventually reached at which the phase relationships necessary for the maintenance of oscillations cease to be preserved. The number of valves that can be used in parallel to increase the power of a transmitter is thus limited.

Another difficulty that arises, with increasing number of valves, is the difficulty of preventing parasitic oscillations. The inductance of the common valve leads, and the valve inter-electrode capacities, and other stray capacities, provide oscillatory circuits of high natural frequency. H/F parasitic oscillations may therefore be set up which rob the main oscillatory circuit of energy.

34. **Push-Pull Transmitting Circuits.**—A push-pull arrangement of valves, as opposed to an arrangement of valves in parallel, necessitates some changes in the relative disposition of the valves and the tuned circuit. These adjustments, however, introduce no new principle, and follow simply from the explanations already given of push-pull amplifier stages, and of transmitting circuits with one valve. A simple push-pull transmitter is shown in Fig. 26 (*a*). The tuned circuit is connected between the anodes of the two valves, and the middle point of the tuned circuit inductance is common with the filaments. This is the ordinary push-pull output circuit as found in amplifiers. The condition for self-oscillations already encountered, that in each valve V_g and V_a must be as nearly in antiphase as possible, has also to be satisfied. Direct grid excitation is used in this transmitter, and so the grid tap and the anode tap of each valve must be on opposite sides of the common filament tap.



PARALLEL FEED.

FIG. 26.

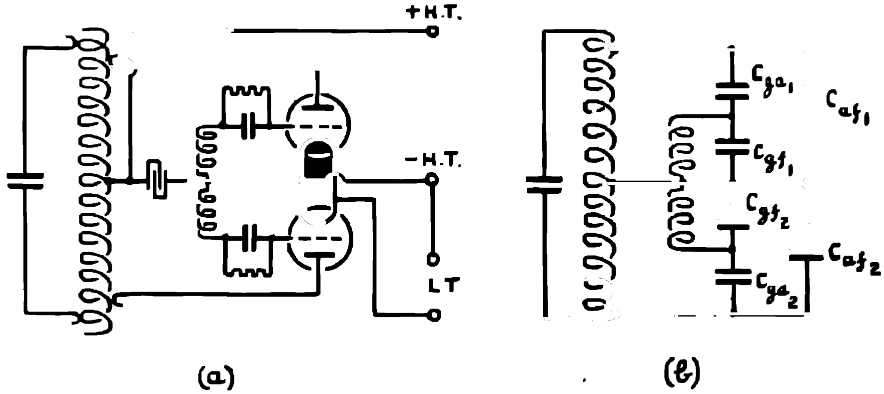
Looking at the disposition of the tuned circuit with regard to either valve, the circuit is seen to be exactly of the type described as—tuned circuit between anode and grid, direct inductive grid excitation. The feed as shown is parallel, and the anode is the electrode of high oscillating potential The oscillatory circuit, as in previous cases, may either be the aerial circuit or be mutually coupled to an aerial circuit. It should be noted that, with a symmetrical circuit of this kind, the physical centre of the tuning inductance will also be approximately the electrical centre, and will be either at "earth" or an "earthy" potential. For this reason it is not always necessary to have a physical connection between the mid-point of the coil F and the filament, in order to achieve the necessary grid excitation. Moreover, the elimination of second harmonic distortion, which is achieved by matching, is facilitated if the centre point of the coil is not joined to the filament.

A push-pull circuit of the type described as tuned circuit between anode and filament, mutual inductive grid excitation, series feed and high oscillating potential anode, is shown in Fig. 27 (*a*). This should be compared with Fig. 14, which shows the same type of circuit with one valve. A large condenser is inserted in the filament lead, as shown, to prevent a short circuit of the H.T. supply.

The use of a grid condenser and leak to increase the efficiency of transmitting circuits, involves, in the push-pull arrangement, that the grid bias of each valve approximates to the condition already described as "curvature biasing," or Class "B" operation. When efficient self-oscillations are being produced, the positive half-cycle of grid voltage in each valve sweeps over the whole of the straight part of the dynamic characteristic of that valve. Thus the circulating current in the

oscillatory circuit corresponds in wave form very nearly to a simple sine curve, and the distortion produced in the wave form with one valve under the same conditions (Fig. 9) is avoided. In other words, the higher harmonic currents, and particularly the second harmonic, which was shown to be principally responsible for this distortion, are almost completely absent. The result is that a much purer wave is radiated, while the higher efficiency of the grid condenser and leak arrangement is preserved.

In designing a transmitter of this nature, having wave changing arrangements, precisely determined conditions of operation are only possible if the grid leak is changed to match the frequency. Unless this is done, the working point of the transmitter may vary over a wide range, since the amplitude of generated oscillations usually varies; at one frequency, conditions may be Class "C" and at another Class "B," or Class "A-B." In all cases, however, if the matching is perfect, there will be no even harmonics although there will be a variety of odd harmonics.



SERIES FEED

FIG. 27.

Another advantage over the parallel arrangement is the diminution of the nett inter-electrode capacity. With two valves in parallel, the corresponding inter-electrode capacities of the individual valves are, of course, also in parallel, and so their effect is doubled. In the push-pull arrangement, it will be seen by inspection of the equivalent tuning diagrams of Figs. 26 (*b*) and 27 (*b*) that corresponding inter-electrode capacities are halved. Changes in these capacities during transmission thus have a much smaller effect on the frequency in the push-pull arrangement. This is more important at high frequencies, where the valve capacities are a larger proportion of the total capacity of the tuned circuit.

It should be noted, that the effective A.C. resistance of two valves in "push-pull" is twice that of a single similar valve.

35. **Peculiarities of Self-Oscillatory H/F Transmitters.**—There is no difference, in principle, between the methods adopted for the generation of high and low radio-frequencies, as should already be evident from the section on H/F receivers, which usually contain self-oscillatory circuits generating H/F oscillations. Thus the classification of transmitting circuits applies equally to all frequencies. The differences that arise are found mainly in the design of components, transmitting valves, etc., required to reach the small LC values of the tuned circuits at high frequencies.

The inter-electrode capacities of a transmitting valve become of increasing importance as the frequency becomes higher. For example, the anode-grid capacity is often used as the tuned circuit

condenser, tuning being effected by varying the inductance. A variation in this capacity produces, therefore, a greater percentage variation in the transmitted frequency than in L/F circuits, and as the same small variation involves an increasing actual change in frequency as the frequency rises, a self-oscillatory H/F transmitter usually produces a wave whose frequency varies between rather wide limits in cycles.

The nature of the variations in inter-electrode capacity have been described in Section B 40, the principal effects being a relatively large *increase* in the grid-filament capacity, and a relatively smaller *decrease* in the grid-anode capacity, during the half cycles when pulses of anode current flow past the grid.

The effect of these variations in the valve capacities may be quite pronounced, more especially since most Service H/F transmitters produce I.C.W. For this reason, all the newer transmitters avoid frequency variation, to some extent, by using circuits designed on the **master control principle** (paragraph 38).

The generation of spurious oscillations also occurs more readily at high frequencies, and provision may need to be made for their suppression. Another important point, when using parallel feed circuits, is the design of the anode choke. The oscillatory frequency at some point of the range, may either be the same as the resonant frequency of the choke and its self-capacity, or a harmonic of it, and standing waves (R.14) may be set up in the choke, of sufficient amplitude to damage its insulation. Further, according as the end of the coke attached to the anode is a node or antinode of potential, *i.e.*, according as the frequency is an even or odd harmonic of the fundamental choke frequency, the oscillatory circuit may either be vitually short-circuited, or a higher P.D. than its rating will allow mav be applied between anode and filament of the valve ; or, when use is made of parallel feed to the anodes of the output valves of a master-controlled transmitter (paragraph 39), the occurrence of choke-resonance producing an antinode of potential at the anodes may render the output stage self-oscillatory. In some cases, Service H/F transmitters are attachments to existing L/F transmitters, and use to some extent the same components. This coke-resonance effect necessitates the use of a different choke for parallel feed H/F and L/F circuits, and, in some cases, the insertion of a large resistance in series with the H/F choke in order to diminish the H/F current when resonance cannot be avoided. Alternatively, provision may be made for short-circuiting portions of the choke, and hence altering its resonances according to the frequency in use.

36. **Particular Self-Oscillatory H/F Transmitters.**—A typical Service transmitter of the older type is shown in Fig. 28 (*a*). The main tuned circuit capacity is the anode-grid inter-electrode capacity of the valves, and a variable condenser is inserted, by means of a switch, either in series or in parallel with this capacity. The two arrangements are shown in Fig. 28 (*b*) and (*c*) respectively. The series position corresponds to the highest frequencies, since the total capacity of two condensers in series is less than that of either. The resistance in series with the anode choke to prevent excessive choke resonance effects should be noted, and also the resistance in the anode lead to damp out spurious oscillations. In the parallel position, these tend to be set up in the circuit comprising the inter-electrode capacity, the blocking condenser, the variable condenser and their leads ; and in both positions they are likely to occur round the circuit consisting of the anode-grid inter-electrode capacities of the valves and their common leads.

The circuit is of the type classified as—tuned circuit between anode and grid, and parallel feed, the anode being the valve electrode at a high oscillating potential. There appears, however, to be no grid excitation, since the connection normally made from the filament to a point on the tuned circuit inductance intermediate between the grid and anode connections is absent. This apparent discrepancy, however, is simply explained by taking account of the valve inter-electrode capacities. At high frequencies, these have values of the same order as the tuning condenser ; in fact, as pointed out above, C_{ga} is the only tuned circuit capacitive path in the series position. The other inter-electrode capacities, C_{af} and C_{gf}, also possess comparable values, and carry a fair proportion of the oscillatory current. The complete tuned circuit, taking account of these capacities, is shown for

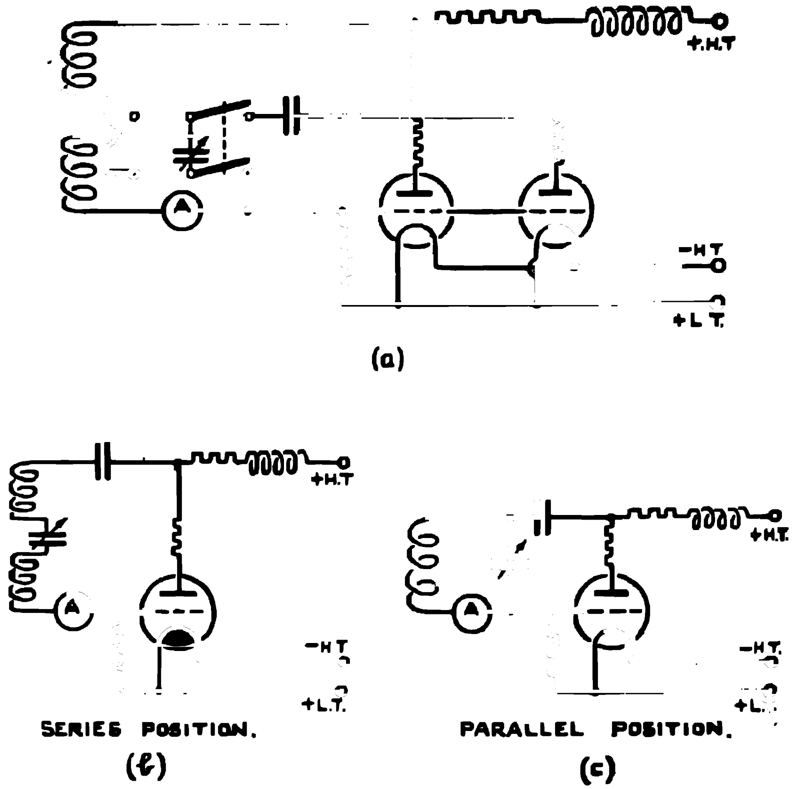


FIG. 28.

the series and parallel positions of the variable condenser in Fig. 29 (*a*) and (*b*). The grid leak resistance is also shown, but for H/F oscillations it is by-passed by C_{gf}. A grid insulating condenser and leak are, of course, still required to give the mean negative grid bias necessary for efficiency. Inspection of these circuit diagrams shows that, in both cases, the grid excitation may be considered as direct capacitive, as in the Colpitts circuit (paragraph 9). The filament connection to the tuned circuit occupies a position in the capacitive branch intermediate between the grid and anode connections. When an oscillatory voltage is developed across the tuned circuit, the filament potential is thus intermediate between the grid and anode potentials, and so V_g and V_a are in antiphase, which is the phasing condition for maintenance of oscillations.

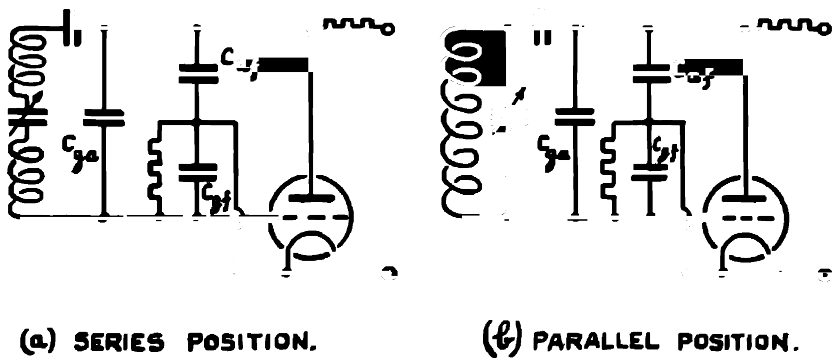
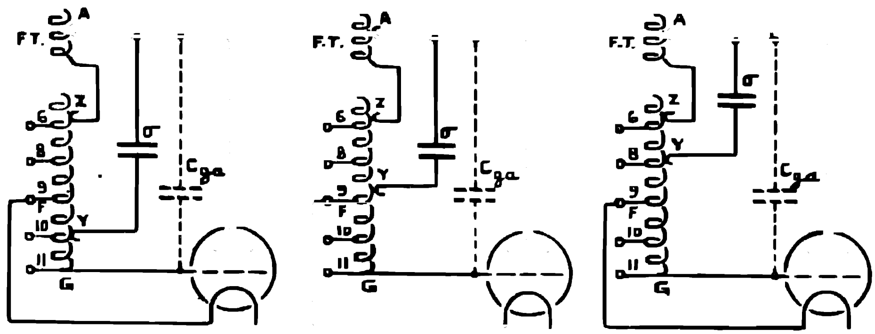


FIG. 29.

Erratum.—Figs.* 28 (*b*) *and* 29 (*a*). *Delete anode blocking condenser shown in top lead of tuned circuit.

An interesting example of the effect of inter-electrode capacity in a transmitting circuit is provided by the Service transmitter already shown in Fig. 19, which was designed to cover the L/F and lower M/F band of frequencies. At the lowest frequencies, tap Y is kept fixed, and tuning is effected by varying taps X and Z. As the frequency increases, the aerial tuning coil is cut out altogether, and the operations of tuning and adjusting the anode tapping point for most efficient conditions are carried out by manipulating the taps Y and Z. In Fig. 30, the equivalent circuits are shown for three positions of tap Y over this range of frequencies. The H.T. supply and smoothing condenser have been omitted, as they do not affect R/F conditions, and so tap A is shown with a direct connection to the anode. This enables the circuits to be exhibited in the simplest way as tuned circuit between anode and grid, direct inductive grid excitation circuits (*cf.* Fig. 5). The lettering exactly corresponds to that in Fig. 19, and should enable the relation between the actual

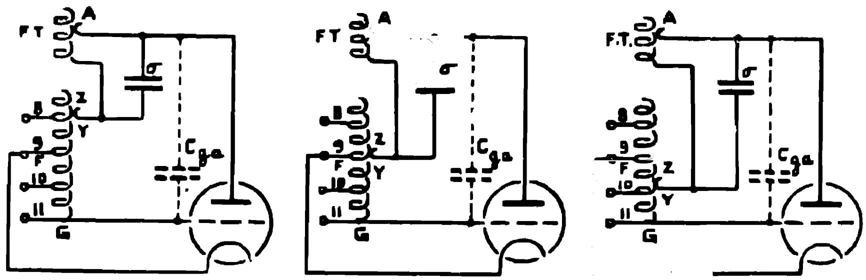


(a) (b) (c)

FIG. 30.

and equivalent circuits to be followed easily. For convenience, tap Z is shown at position 6 on the tuned circuit inductance in each case, but it may occupy any lower-numbered position than tap Y. It will be seen that the current flowing through C_{ga} becomes of increasing importance as the frequency increases ; C_{ga}, in fact, forms the capacitive branch of the tuned circuit in Fig. 30 (*b*) and (*c*), and enables the grid excitation (the oscillatory P.D. from 9 to 11 of the inductance) to remain in the correct phase for maintenance of oscillations ; the parallel circuit consisting of σ, and the part of the oscillatory circuit inductance from A to Y, is resonant at a higher frequency than that on which the whole circuit actually oscillates, and so has a nett inductive reactance at the oscillatory frequency. Even at the lower frequency corresponding to Fig. 30 (*a*), C_{ga} plays an important part in supplying the grid excitation, although, as the filament connection is then between the anode tap A and tap Y, grid excitation in the correct phase is provided by the P.D. from 9 to 10 of the oscillatory inductance, while still considering the aerial capacity σ as part of the capacitive branch of the tuned circuit.

It was found that considerably higher frequencies could be obtained with this transmitter by combining taps Y and Z at a common point on the oscillatory inductance. The equivalent circuits for three such positions are shown in Fig. 31 (*a*), (*b*) and (*c*). With the same amount of fine-tuning inductance (F.T.) in circuit, the frequency increases as the combined tap ZY is moved from 8 to 10 on the oscillatory inductance. In each case, the aerial capacity (which includes the H/F condenser, 0·25 jar), and the fine-tuning inductance form a parallel circuit in the inductive branch of the main tuned circuit ; this parallel circuit always tunes at a higher frequency than the whole circuit, and so possesses inductive reactance at the actual frequency of oscillations, *i.e.*, it can effectively be replaced by an inductance. In Fig.31 (*a*), oscillations can still be obtained with the whole of the fine tuning coil out of circuit, since the branch of the main tuned circuit from A to F, through the oscillatory inductance from 8 to 9, possesses inductive reactance (σ is short-circuited).



(a) (b) (c)

FIG. 31.

This is not the case in the circuits of Fig. 31 (*b*) and (*c*), and part of the fine tuning coil must then be included in the circuit for oscillations to be possible

In all these cases of Fig. 31, there is a considerable departure from maximum efficiency, and the amount of power transferred from the valve to the oscillatory circuit is only a small proportion of the power input. The power dissipation at the valve anode is correspondingly increased, and care must be taken to prevent this from rising above the anode rating, and hence damaging the valve. This is particularly necessary for the adjustment shown in Fig. 31 (*c*), where the only part of the grid excitation in the correct phase is the P.D. from points 10 to 11 of the oscillatory inductance. The whole anode (make-up) current flows through the inductance from 9 to 10, and so produces a P.D. which is 90° out of phase with that from 10 to 11. V_a is thus by no means 180° out of phase with V_g at this adjustment, and the efficiency is low. In some cases, it is not possible for oscillations to be generated at all at this setting, one determining factor being the value of the aerial capacity.

37. **Frequency Variation.**—In the elementary treatment of the valve as a generator, given in the earlier paragraphs of this section, the frequency of the self-oscillations was taken as the natural frequency of the oscillatory circuit, $f = \frac{3 \times 10^4}{2\pi\sqrt{LC}}$, where L is in mics., and C in jars.

This is only a first approximation to the truth, and a more exact analysis shows that the frequency depends, in addition, on the valve constants r_a and g_m; it also depends on the LC values of all circuits coupled to the oscillatory circuit, such as the aerial circuit, and that formed across the inter-electrode capacities of the valves. Unless all of these components remain constant during oscillatory conditions, the frequency will inevitably vary.

In practice, there are very many reason why changes take place in values of the various components determining the generated frequency. The valve constants r_a and g_m can only be regarded as at all constant over the straight parts of the mutual characteristics, and even then their values are slightly different for differing steady potentials of the anode and grid. When the tuned circuit is between anode and grid, the anode-grid inter-electrode capacity is in parallel with the tuning condenser. Now inter-electrode capacity depends, partly, on the space charge and electron density in the valve (paragraph 35), *i.e.*, on the P.Ds. between the electrodes, and on the filament emission, and changes in either of these quantities alter the values of the various inter-electrode capacities. For both of these reasons, it follows that frequency constancy cannot be achieved unless the steady potentials of the anode and grid, and the filament emission, are kept constant. In practice some fluctuation is inevitable, and for this and other reasons the frequency of a valve generator may vary correspondingly.

As an indication of the change of frequency to be expected, due to a change of anode voltage and filament voltage, it may be observed that in the case of a paticular valve self-oscillating at a radio-frequency of 3,000 kc./s., the frequency changes by 50 cycles for a 10 per cent. change of

anode voltage, and by 1,600 cycles for a 10 per cent. change of filament voltage; these values are only approximate, and vary with the varying stiffness of the circuit.

For many reasons a limit must be set to the amount of permissible frequency variation. If the variation is at all considerable, the heterodyne reception of C.W. becomes difficult; this is particularly the case at high frequencies, where the heterodyne note will be constantly changing in pitch. Frequency variation of an I.C.W. signal will give at least three notes, two rising and one falling in pitch, as the carrier frequency varies. Apart from these difficulties, the limited number of lines of communication in the æther spectrum, together with the steadily increasing volume of W/T traffic, make it essential that each transmitter should adhere to its allotted frequency within very narrow limits.

The allocation of frequency bands, for all purposes, and the determination of the frequency tolerance of the transmitters, forms part of the work of the Comité Consultatif International Radio-éléctrique. The following table indicates the recommended frequency tolerances for certain types of station. The tolerance is expressed as a percentage, and is merely the permissible drift during transmission.

Frequency range in kc./s.	Recommended frequency tolerance.	
	Land stations.	Mobile stations.
	Per cent.	Per cent.
10–550	0·1	0·5
550–1500	0·1	0·5
1500–6000	0·1	0·1
6000–23000	0·1	0·1

In the Naval organisation, a much more stringent requirement is frequently superimposed, namely, that a transmitter tuned to any specified frequency must, at all times, and without further tuning, oscillate at that frequency within a few hundred cycles.

It is possible to recognise three varieties of frequency variation; these will be referred to under the headings · (*a*) frequency drift; (*b*) slow frequency variation; and (*c*) frequency spread.

It will be shown that the various causes may be classified into disturbances originating in the circuit, and disturbances originating in the valves.

(*a*) FREQUENCY DRIFT.—Frequency drift is chiefly due to the expansion of the tuning coil by heating. The inductance is thereby increased, thus causing a decrease in transmitted frequency. The resistance of the coil also increases with temperature and, therefore, tends further to decrease the frequency; this effect is only of secondary importance compared with that due to increasing inductance. In practice, this has been countered by the design of special "anti-drift" coils.

(*b*) SLOW FREQUENCY VARIATION.—The slow frequency rise and fall of self-oscillatory circuits, in which the aerial comprises the only tuning capacity, is the result of aerial sway. In many ships, the aerial feeder runs parallel and adjacent to a steel mainmast, and provides a large proportion of the total aerial capacity. The sway of the feeder in the wind, or with the roll of the ship, varies its position relative to the mast, and therefore its capacity. A change of weather conditions may alter the length of the guy ropes locating the feeder, and provide the reason why the transmitter will not always return to the same frequency with the same coil adjustments.

(*c*) FREQUENCY SPREAD.—Fluctuations of the anode, grid, and filament voltages are reflected in changes of the valve constants. Several of the changes which occur, produce opposing

effects on the generated frequency, but, on the average, probably the most potent cause of variation is the increase of grid/filament capacity with increase of electron density in the field between the electrodes. The pulses of anode " charging current " at each positive half cycle thus introduce a frequency variation (frequency modulation) ; the frequency is therefore continuously changing over a narrow band.

Now this effect is more noticeable in H/F circuits where the valve inter-electrode capacity becomes comparable with that of the tuning condenser. Moreover, if such circuits are over-loaded or badly adjusted, secondary emission from the grid may occur, particularly when using grid leak bias. The mean grid potential then fluctuates, and may introduce an amplitude modulation.

In a ship installation, vibration is a frequent cause of trouble. In the silica valves, it causes a small and continuous change of the relative positions of the grid and filament, the result being a continuous small variation in the valve characteristics which may modulate the transmitted wave at noise frequencies. The deciding factor is the " working point " of the valve. If the bias is beyond the cut-off point, as in Class " C " operation, a grid modulation of small peak amplitude can be applied to the grid without affecting the anode current, and therefore the aerial current. If the bias point is somewhere on the bend, as it often is in H/F transmitters, any modulation applied to the grid will cause a variation in the anode current, and a modulated wave will be radiated. Any change in the position of the anode within the valve, produces an effect of much smaller importance.

The general effect of all these modulations, produced by disturbances originating in the valves, is to produce the appearance of a "frequency spread." At the receiver, the tuning will appear to be flat ; heterodyne reception of such a transmitter will produce a gruff note, having no definite musical quality. The transmission is therefore difficult to read through atmospherics, although under good conditions, the gruff note is less tiring to the ear for long periods of working, and has what the receiving operator terms " punch."

If any one of these forms of frequency variation is present in exaggerated form, interference will be caused to other services on adjacent frequencies.

38. **Methods of Avoiding Frequency Variation.**—It has been seen that pure C.W. can only be generated by a self-oscillating valve circuit, when both filament and anode supplies are from a source free from any modulating influence, when the valve is carefully insulated from any mechanical vibration, and when the oscillatory circuit itself is entirely rigid in design. In general, it is only possible to satisfy these conditions with comparatively small transmitters, of a few watts input ; for larger powers, other methods must be employed for stabilising the frequency. It is not over-stating the case to say that the high powered self-oscillatory transmitter cannot produce a frequency sufficiently constant for modern requirements.

In modern practice, there are three general methods of generating a frequency which is stabilised within the prescribed limits of frequency tolerance ; they will be referred to under the headings :—

(*a*) MASTER OSCILLATOR CONTROL.—This will be taken to include all cases where the self-oscillatory MASTER CIRCUIT gives a frequency which may be continuously varied over a band ; in all cases it will be a low power rigidly designed circuit, which passes its output on to a power amplifier, and thence to the aerial. In the Service, such circuits may give a frequency constancy of the order of $\pm$ one part in 2,000, and are described in further detail below. With lower power, and more stages, a good circuit should have an error of $\pm$ 1/5,000.

(*b*) TUNING FORK OR QUARTZ CONTROL.—These are electro-mechanical methods for the MASTER CONTROL of frequency ; only the latter method is seen in Service practice. It is described in further detail below, and by means of suitable circuits it is easily possible to produce pure C.W. with a frequency constancy of the order of $\pm$ one part in 10,000. This method of MASTER CONTROL does not provide a continuously variable range of controlled frequencies ; it is generally used to give a limited number of precisely determined " spot frequencies."

(c) SPECIAL FREQUENCY STABILISED CIRCUITS.—Very many self-oscillatory circuits have been designed to generate a frequency which is substantially independent of one or the other of the various variable factors. For example, there are circuits in which the frequency produced is independent of any change in the valve constants r_a and g_m, and which minimise the effects of any alteration in the inter-electrode capacities. These circuits give a measure of frequency control, but are not as efficacious as either of the two foregoing methods, which should therefore be preferred whenever possible. A brief account of some of these circuits will be given below.

The term MASTER CONTROL will be used in reference to both methods (a) and (b); the inception of transmitters with a stabilised drive circuit, followed by a power amplifier, opened a new era in valve transmitters (paragraph 1).

39. **Master Oscillator Control.**—The general principle of the method adopted, is to generate the self-oscillations in a circuit the constants of which are sufficiently under control that the frequency generated varies within very narrow limits. The oscillatory voltage produced at this frequency is then applied between grid and filament of a valve or valves, capable of giving sufficient output power for the transmission required. The oscillations in the power output stage are thus forced oscillations, and not free oscillations, as in the self-oscillatory circuits previously considered, and they take place at a frequency entirely decided by the master oscillator. This power stage is thus essentially the same in principle as the power amplification stage in a receiving circuit, though the power output is, of course, enormously greater. At Rugby, the output of one master circuit is 10^{-11} times the aerial power. The circuit in which self-oscillations are initially generated is called the MASTER OSCILLATOR, the reason for the name being obvious.

An analogy is sometimes drawn between the balance wheel of a watch and the master oscillator part of a transmitter. The balance wheel allows the main spring to administer the regular impulses to the watch mechanism, at a frequency entirely determined by the balanced wheel and associated hair spring. The parallel is an interesting one, but it cannot always be drawn too closely.

The simplest type of master oscillator is merely a rigid low-powered self-oscillatory circuit, in which special precautions are taken to avoid variation in the electrical quantities which affect the frequency. Particular attention is paid to such points as the smoothing of the H.T. supply voltage to the anode, the careful regulation of the filament supply, the position of the working point on the valve characteristic determined by the grid leak and condenser combination, the use of valves which need not be worked up to the limit of their anode rating, and by the use of a tuned circuit in which the tuning capacity is relatively large in order that variations of inter-electrode capacity may affect the frequency as little as possible. "Frequency drift" is countered by the use of anti-drift inductance coils, and other components rigid in design. The whole circuit should be mounted in such a way as to prevent mechanical vibration of its parts; in one Service case, the silica valves were tilted at a small angle to the vertical, to assist in limiting the internal vibration of the electrodes and their supports.

These requirements are more easily fulfilled at low frequencies, and, in general, it may be said that master circuits require more careful design as the frequency increases.

Master circuits must be carefully screened from the main power stage, and the components of both circuits arranged so as to minimise stray coupling. Inter-electrode capacity coupling between the output of the power amplifier and the master oscillator must also be carefully neutralised. Using the analogy of the watch, there must be no reaction of the main spring on the hair spring. This point is dealt with in more detail below. For the above reason also, the use of any properly adjusted master circuit ensures that no power is transferred from the master circuit to the aerial circuit, and *vice versa*; thus the effects of variations in the aerial capacity and inductance on the frequency are eliminated.

A transmitting circuit of this type, designed for medium frequencies, is shown in Fig. 32. The master oscillatory circuit is of the type described as—tuned circuit between anode and grid, direct inductive grid excitation, parallel feed, the anode being the electrode at a high oscillatory potential. The main circuit consists of two valves in parallel, their external anode circuit being the aerial circuit. Parallel feed is also employed for the main power valves. The oscillatory P.D. developed across the portion of the master tuned circuit inductance between the grid and filament taps, is applied to the grids of the power valves through a coupling condenser C_1. N.C. is the neutralising condenser, to balance out inter-electrode capacity coupling between main and master circuits.

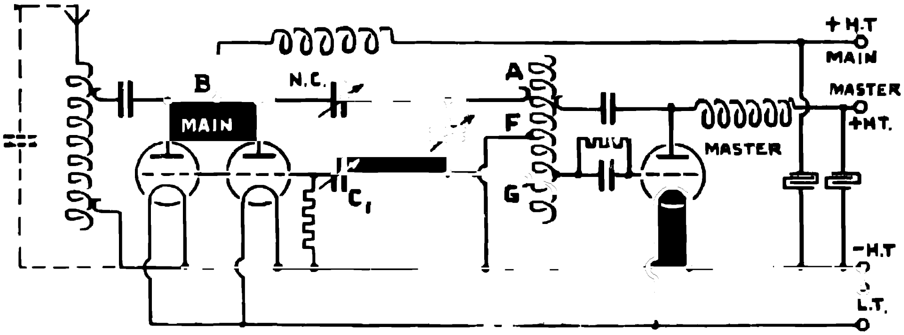


FIG. 32.

The necessity, in general, for neutralising master controlled transmitters should be carefully noted. It will be recalled that, in the case of receivers, the necessity for neutralising H/F amplifying stages passed with the invention of the screen grid valve, and, later, the H/F pentode. That change in W/T receiver construction dates from about 1929, in which year the technical press contained articles with titles such as "The passing of the neutralised triode." A similar change has begun to affect transmitter construction.

The best master controlled transmitters ought to include a buffer stage between the master circuit and the power amplifier, the grids of the buffer valves being biased so far back that they take no grid current, and therefore apply no load to the master. Space limitations, and the necessity for rapid wave changing, usually make it impossible to use ideal designs in H.M. Ships.

40. **Neutralisation of Master Oscillator Circuits.**—The difficulties that arise in securing adequate prevention of regenerative reaction between the main and master circuits present a complex problem, and only the elements of it can be considered here. Unless such reaction is eliminated, the transmitted frequency may vary even more than would be the case with a self-oscillatory circuit. Any reaction of the main on the master circuit renders the frequency of the master dependent, to some extent, on the constants of the main circuit, and re-introduces in more virulent form the frequency variations which the use of a master oscillator is designed to avoid. Conditions may also occur in which both the main and master circuits are rendered self-oscillatory. As the natural frequencies of the two circuits are never likely to coincide for any period of time, the effect of self-oscillatory conditions in both circuits, is that the oscillations in the main circuit are modulated at the frequency of the master. Three frequencies will then be transmitted instead of one, all three in addition being variable. (This point will be better understood after the explanation of side-bands in Section "N.") The necessity for careful screening has already been emphasised, and it is also necessary to neutralise the flow of energy through the coupling condenser and the grid-anode inter-electrode capacities of the power valves. As a typical instance, the neutralisation of the circuit shown in Fig. 32 will be discussed. A neutralising condenser is shown connected between the points B and A. The operation of such condensers has already been explained for amplifying

circuits. Fig 33 shows one aspect of the neutralisation of this circuit, the relevant parts of Fig. 32 being redrawn as an A.C. bridge with respect to feed of energy from the master, and may be compared with the "tapped input" method of neutralising (Section "F").

It is obviously of the greatest importance, in the operation of such a master oscillator circuit, to ensure accurate neutralisation, and for that reason the purpose to be kept in mind when adjusting the neutralising condenser will now be treated in greater detail.

Suppose that the master circuit is oscillating, and that the filaments of the main power valves are alight, but that their H.T. supply is not made ; also that the coupling condenser is set at a fixed adjustment. In Fig. 33 there is then an oscillatory P.D. between A and G, and oscillatory current will flow through the two parallel paths presented by the master inductance, AFG, and the neutralising condenser N.C., anode-grid valve capacities and coupling condenser, ABG. One aim of neutralisation is to adjust N.C. so that the points B and F are at the same oscillatory potential as far as this applied P.D. for the master is concerned. If this is ensured, there can then be no oscillatory P.D. across BF from this cause, and therefore no energy can be FED FORWARD from the master

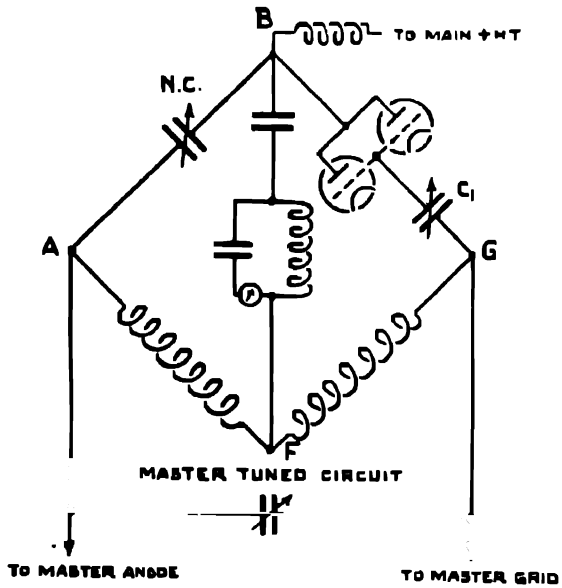


FIG. 33.

circuit to the output circuit. Hence, the condition for neutralisation is that the voltage drop across AB should be equal in magnitude and in phase to that across AF (necessitating also the equality of the voltage drops across FG and BG). The neutralising condenser must be adjusted until this condition is satisfied.

Of more importance, however, is the question of FEED-BACK OF ENERGY from the main tuned circuit to the master inductance when the main H.T. supply is made. The reactance of the large coupling condenser C_1 is small, and so the main circuit (across BF),in series with the master inductance from F to G, is effectively in parallel with the valve capacities between the anodes and grids. This constitutes a possibly self-oscillatory circuit of the type classified as tuned circuit between anode and grid and direct inductive grid excitation, the latter being across the inductance FG. The introduction of the neutralising condenser enables an equal and opposite P.D. to be developed across the other part, AF, of the master inductance ; the bridge becomes balanced, and no P.D. can be produced across AG.

It can easily be seen that the setting of the neutralising condenser to prevent "feed back" of energy is the same as that demanded by the conditions for no "feed forward." With this setting, therefore, self-oscillation of the main circuit is prevented, and also any reaction of main on master

circuit, and *vice versa*. The condition for "no feed forward" usually provides the practical method of neutralising.

To obtain the correct adjustment, the following procedure may be adopted, the conditions being as postulated at the beginning of the discussion. When the adjustment is incorrect, there will be an oscillatory P.D. across the main circuit. This circuit is tuned until its ammeter registers maximum current. The neutralising condenser is then adjusted until the ammeter reading falls to zero.

It will be observed that the arm ABG of the bridge is in parallel with the master oscillatory inductance, and so any alteration of either the neutralising condenser or the coupling condenser alters the master tuning. Hence, when neutralising, the main circuit must be re-tuned to give maximum ammeter reading at intervals during the adjustment. When approximate neutralisation has been obtained, the final frequency adjustments of the master and main circuits can be made, and the process repeated.

A factor rendering good neutralisation difficult, is the inductance of the leads in the bridge arms containing the neutralising condenser, the power valves and the coupling condenser. When this is appreciable, the setting of the neutralising condenser varies with the frequency ; this is especially the case at H/F.

A common disturbing factor is the occurrence of H/F parasitic oscillations in the circuit formed by the inter-connection of the valve anodes and grids, as has already been mentioned when considering valves in parallel. Another circuit which may generate parasitic oscillations is provided by the master tuning condenser, and the arm ABG of the bridge in conjunction with the connecting leads. These oscillations may be damped out by inserting resistances in the arms of the bridge. To preserve phase balance in the bridge, the resistances must be adjusted symmetrically in each limb. This introduces damping into the main and master circuits.

Another method is to eliminate the grid excitation of the valves at the parasitic frequency, by inserting, between the grid and its leak resistance, a coil and condenser in parallel, of reactance equal and opposite to the original grid filament reactance at this frequency.

There is then no grid-filament P.D. at the parasitic frequency, and so these spurious oscillations cannot be set up. The use of this method requires an alteration in the setting of the neutralising condenser, whenever a chance is made in the frequency it is desired to transmit.

Alternatively, it may be considered that this introduction of a coil and condenser in parallel alters the parasitic frequency so as to make it an odd harmonic of the anode choke resonant frequency. The standing waves then set up in the choke have approximate voltage nodes at each end, and so there is practically no anode voltage variation at the parasitic frequency, *i.e.*, self oscillations cannot be maintained.

41. **Master Controlled H/F Transmitters.**—Reference has been made to the greater frequency variation to be expected in H/F self-oscillatory circuits ; this is particularly the case when I.C.W. is being produced. Modern self-oscillatory H/F transmitters are capable of producing C.W. with very little frequency modulation, but I.C.W. is commonly used for H/F transmission, since it is equivalent to three C.W. oscillations at neighbouring frequencies, and is therefore of some assistance in overcoming fading (P.14). To keep the carrier frequency constant, some form of master control is desirable.

A master control circuit, using a push-pull output stage, is shown in Fig. 34.

The master is a rigid low-power self-oscillatory circuit, of the type classified as tuned circuit between anode and grid, direct inductive grid excitation and series feed. The output across the master tuned circuit inductance is taken to the grids of the power valves through two coupling condensers. Parallel feed is used for the H.T. supply to the anodes in the power stage, and provision is made for avoiding choke resonances by short circuiting part of the anode chokes according to the frequency in use.

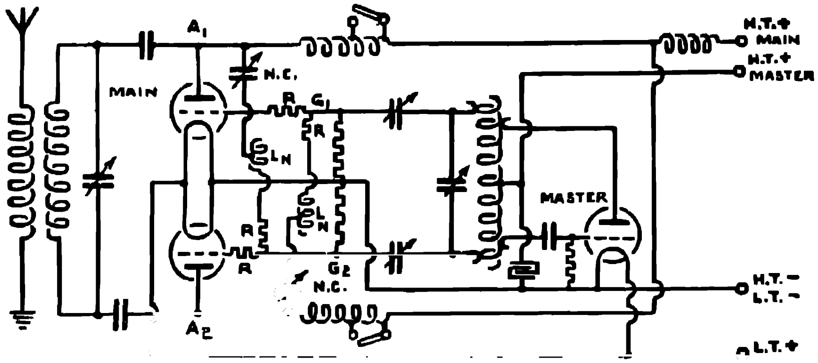


FIG. 34.

This circuit provides an interesting example of the principles of neutralisation. Neutralising condensers are connected between the anode of each valve and the grid of the other, but their introduction tends to give rise to parasitic oscillations round the neutralising bridge circuit, Fig. 35 These are damped out by the symmetrical insertion of resistances in the four arms of the bridge, and the bridge is then readily balanced for frequencies at which the inductive reactance of connecting leads can be neglected, *i.e.*, an applied P.D. between A_1 and A_2 gives rise to no P.D. across G_1G_2, the master circuit.

Although such a balance obviates reaction on the master from the main circuit, it does not necessarily prevent self-oscillation of the latter. The filaments of the valves are earthed, and, at any moment, the anodes A_1 and A_2 have equal and opposite potentials to earth. Thus, although G_1 and G_2 must be at the same potential when the bridge is balanced, they are not necessarily at earth potential, *i.e.*, there may be a grid-filament P.D. to earth in each valve. Inspection of Fig. 35 shows that this grid excitation is in phase with the anode-filament P.D in one valve, but in anti-phase with it in the other. This causes one valve to absorb energy, but the other will tend to set up self-oscillations in the main circuit if the grid excitation is sufficient. Hence it is further necessary to balance the bridge in such a way that G_1 and G_2 are at earth potential, *i.e*, midway between the potentials of A_1 and A_2. Assuming the valves to be matched, so that their inter-electrode capacities

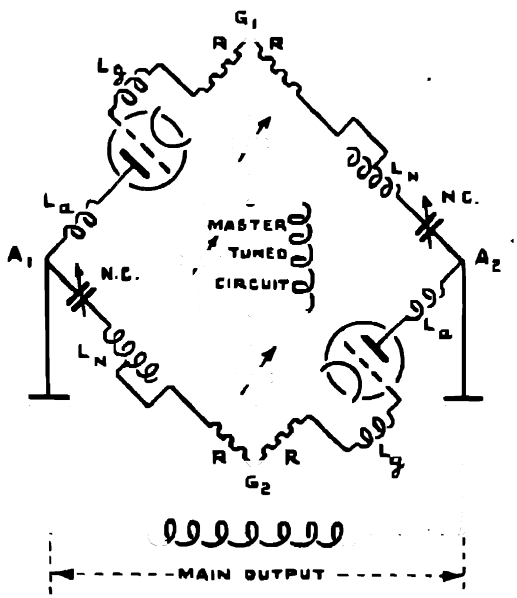


FIG. 35.

are equal, this further condition is satisfied if the bridge is balanced with equal settings of the two neutralising condensers, and for this purpose the condensers are ganged.

Above a certain frequency, the inductive reactance of connecting leads is large enough to make a balanced bridge impossible, unless the inductance is symmetrically disposed in the four arms. In Fig. 35, the inductances of the leads from the anodes and grids of the valves are shown as L_a and L_g respectively The condition for a balance which is independent of frequency, and brings G_1 and G_2 to earth potential, is that the capacity, inductance and resistance of each arm of the bridge should be the same. The low frequency balance ensures that the capacities are equal, and the anti-parasitic resistances are symmetrically inserted, but it may be necessary to introduce small coils into certain of the bridge arms, to give a symmetrical distribution of inductance. This is indicated by the variable coils L_N in Figs. 34 and 35.

Neutralisation is then obtained by a series of successive approximations, as follows :—

(*a*) Set each inductance L_N, by calculation, to a value approximately equal to $L_a + L_g$.

(*b*) Energise the circuit at the lowest frequency to be transmitted, and adjust the ganged neutralising condensers until a balance is obtained. (If the neutralising condensers are not ganged, they must be adjusted equally).

(*c*) Energise the circuit at the highest frequency to be transmitted, and re-neutralise by equal adjustments of the inductances L_N

(*d*) Repeat operation (*b*).

This process can be carried out when the transmitter is first assembled, and should give approximately correct neutralisation over the required frequency range ; but if the settings are found to be critical, slight adjustments may be necessary whenever the frequency is altered.

42. **Tuning Fork or Quartz Control.**—The other type of master oscillator circuit is one in which an endeavour is made to render the frequency of the self-oscillations independent of the electrical constants of the circuit. The control may be exercised, for example, by a tuning fork vibrating at a definite frequency.

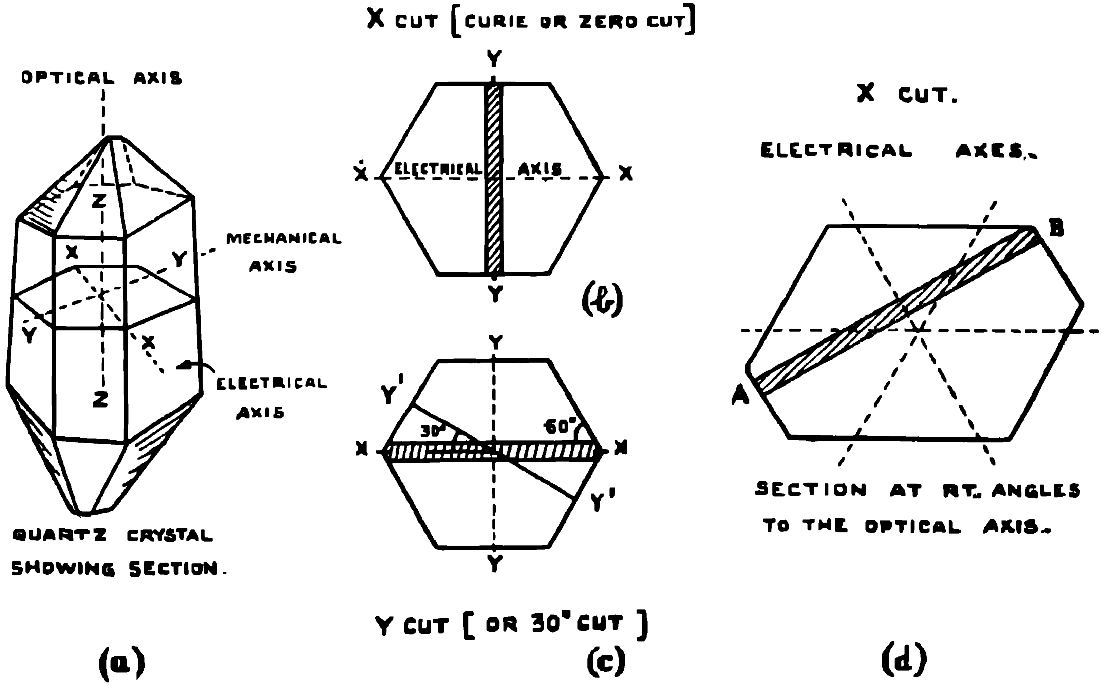


FIG. 36.

The commonest master oscillator of this nature makes use of the special properties of certain crystalline materials, notably quartz, and is therefore called QUARTZ CONTROL. Other materials which have been used are Tourmaline, and Rochelle salt.

QUARTZ occurs in nature, and often in large lumps having the crystalline form shown approximately in Fig. 36 (*a*). It is particularly abundant in Madagascar and Brazil. Chemically, it is an oxide of silicon (SiO_2), called silica, and silica is merely fused quartz. The crystalline structure is of the type technically known as "hemihedral with inclined faces." Its optical behaviour determines a particular direction in the crystal called the optical axis; this direction is that of any line in the crystal parallel to the line joining its two extremities.

The peculiar property of such a crystal on which quartz control depends, is that if it is subjected to pressure, an electrical P.D. is developed across it. The application of a tension instead of a pressure produces a P.D. in the opposite direction. Conversely, if a P.D. is applied across the crystal, it undergoes mechanical deformation, *i.e.*, it contracts or expands according to the direction of the P.D. It is found that the relation between pressure and P.D. follows almost a linear law.

These properties were discovered and investigated by the Curies in 1880 (*cf.* K.1). The effects are said to be PIEZO-ELECTRIC (Gk. PIEZEIN—to Press) and their magnitude varies according to the direction in the crystal along which the pressure or P.D. is applied.

Investigation has led to the discovery and definition of further axes of symmetry, namely, ELECTRICAL AXES, and MECHANICAL AXES. Quartz crystals are seldom symmetrical in form but for simplicity, in Fig. 36 (*b*) and (*c*), the sections perpendicular to the optical axes have been taken to be hexagonal in shape.

ELECTRICAL AXES.—These are three in number and are directions which are defined by lines bisecting the angles of the hexagon which is formed by the sections perpendicular to the optical axis; Fig. 36 (*a*) and (*b*) shows one of these XX, and others for an irregular shaped crystal are shown in Fig. 36 (*d*). The electrical axes are directions of maximum piezo-electric activity.

MECHANICAL AXES.—These are also three in number and are directions defined by lines drawn perpendicular to the sides of the hexagon formed by the section perpendicular to the optical axis. Fig. 36 (*a*) and (*b*) shows one of these axes YY, and a section Y'Y' is shown in Fig. 36 (*c*). They are the directions of mechanical stress corresponding to the electrostatic stress.

Crystal sections are used in the form of plates or bars. There are very many different sections and each is defined by the angle at which it is cut. The best known are still, probably, the X-cut and the Y-cut crystals.

X-CUT.—This section is also known as the "perpendicular cut," since its major surfaces are parallel to the optical axis and perpendicular to an electrical axis, as shown in Fig. 36 (*b*). X-cut crystals develop a large piezo-electric effect, and they execute longitudinal vibrations; an expansion or contraction along the mechanical axis is accompanied by a contraction or expansion along the electrical axis respectively. These crystals have two possible frequencies of oscillation, and the one chosen depends upon the working conditions. The temperature coefficient is negative, and has a value of the order of 20 parts in 1,000,000 per degree Centigrade. Moreover, as the temperature alters, there are usually discontinuities in the frequency/temperature coefficient curve; this produces an abrupt jump in the frequency, and constitutes a source of trouble. This cut, or one of the newer ones, is used in high-class working.

Y-CUT.—This section is also known as the "parallel" or 30° cut, because of the 30° change in orientation from the perpendicular; it is shown diagrammatically in Fig. 36 (*c*). It is not so commonly used as the X-cut, but the piezo-electric output is usually greater; very frequently the temperature coefficient is positive and of the same order of magnitude as in the previous case. Y-cut crystals are usually subject to more discontinuities than X-cuts; they are often used in small sets where maximum crystal output is needed.

Very many other " cuts " of crystal have been investigated, often with the object of discovering one having as zero temperature coefficient. Very satisfactory sections have recently been obtained by cutting plates at different angles of rotation about the X-axis, *i.e.*, with the plane of the plate making some angle less than 90° with the optical axis. " Zero temperature cut " crystals are in use in the installation on board R.M S. " Queen Mary," and also in the Service.

A crystal usually executes either longitudinal or shear vibrations, but if long rods or bar are used instead of crystal plates, it is possible to produce a torsional (or twisting) vibration, or flexural (or bending) vibrations. The subject is a complex one, and the frequency will vary with the mode of oscillation.

In the case of X-cut crystals performing longitudinal vibrations, the natural frequency varies inversely with the linear dimension principally concerned ; it is given approximately by the expression—

$$f = \frac{K}{t}$$

where $K = 2 \cdot 860 \times 10^6$, and where t is the thickness in millimeters If the slice is one millimeter thick, it will be seen that the frequency generated is approximately 3,000 kc./s. X-cut crystals can also oscillate at another frequency, given by the above expression in which the width replaces the thickness " t." It will be seen that this is a lower frequency and, in practice, adjustments are made to use the higher one only These crystals can be ground having natural frequencies ranging from about 25 kc./s. to about 7,500 kc/s. ; the practical upper limit is more usually of the order of 4,000 kc./s., for at frequencies above this the crystal is very thin and is therefore not very robust. Recent improvements in the technique of crystal grinding have made it possible to produce crystals capable of use, under certain conditions, at frequencies as high as about 15,000 kc./s.

MECHANICAL RESONANCE.—Suppose now that an oscillatory P.D. is applied across such a crystal slice. This sets up an alternating stress in the plate and vibration commences. As in the electrical case of an alternating E M.F. applied to the oscillatory circuit, this vibration may be considered to consist of two components, a forced oscillation at the frequency of the applied P.D., and a free oscillation at the natural frequency of the plate. The frequency of the free oscillation, depends only on the mechanical properties and the dimensions of the plate, and not at all on that of the applied P.D. In general, the forced oscillation is of small amplitude, but if the frequency of the applied P.D. is anywhere near the natural frequency of the crystal, conditions approaching mechanical resonance ensue, and the magnitude of the free oscillation reaches a large value. This free vibration sets up an oscillatory P.D. of correspondingly large amount across the crystal faces. It is the latter oscillatory P.D. which may be applied between the grid and filament of a valve to maintain electrical oscillations in the valve and its associated circuit. The frequency of such oscillations is found to be almost the natural mechanical frequency of vibration of the quartz plate, and is almost independent of the electrical properties of the valve circuit. Frequency variation, due to changes in these properties, is largely eliminated and the transmitted frequency is constant.

Since any crystal has a natural frequency of oscillation, it is reasonable to suppose that it may be replaced by an equivalent LC circuit ; the behaviour at resonance indicates that such a circuit would be highly selective, and therefore with a very small " log dec." This conception is a very practical one, and its use assists materially in the appreciation of the action of quartz control circuits.

Although other crystals possess piezo-electric properties, quartz is still the most commonly used. Tourmaline is a semi-precious stone and is therefore too expensive. Rochelle salt is finding extensive application in connection with gramophone pick-ups and microphones, particularly in view of the fact that it is more sensitive than quartz and can be artificially manufactured. It is, however, very fragile and suffers from a number of other defects which, at present, prevent it from being so extensively used as quartz.

43. **Mounting the Quartz Crystal.**—The two factors which affect the frequency at which the crystal oscillates, apart from its mechanical dimensions, are temperature and the size of the air gap in which it works. Depending on the way in which the crystal is cut, the temperature coefficient

may be positive, negative, or almost zero. Crystals are usually operated in horizontal mountings, so that, in operation, they "float" between parallel metal surfaces which give an air gap, the thickness of which is the same at all points. If crystals are mounted vertically they tend to lean on one plate or the other, thus producing a tapering air gap at each side ; the thickness of the gap at some point may then be such as to set up air vibrations which damp those of the quartz, and, in practice, the oscillations are difficult to maintain.

Fig. 37 represents a modern type of crystal holder in which the crystal is completely enclosed. Essentially, the holder consists of two large masses of metal separated by insulation, the crystal resting on the lower one. The two metal faces between which the crystal is held are accurately plane and parallel, and a very delicate adjustment of the air gap between the upper face and the crystal may be made by means of the micrometer screw down in the diagram.

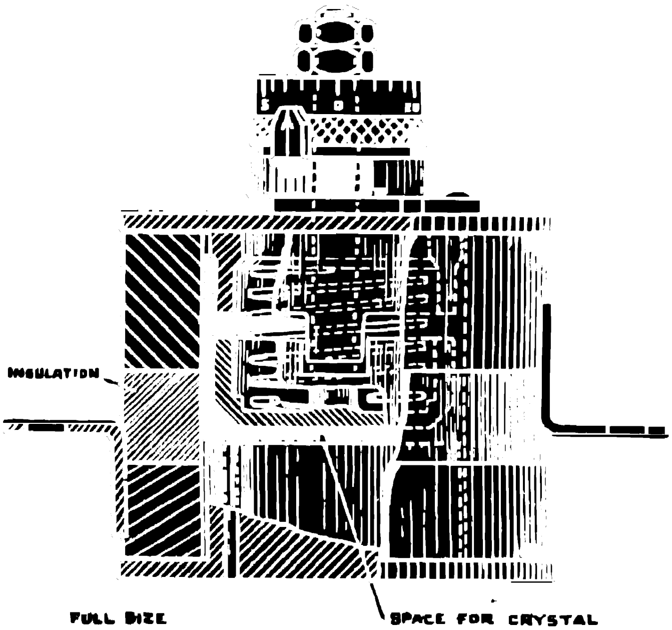


FIG. 37.

The frequency at which the crystal vibrates varies with the thickness of the air gap, due to the change in the effective capacity of the crystal holder plate system ; the frequency becomes higher with a larger gap—or small capacity holder—and lower with a smaller gap, provided that the crystal does not jump to another mode of vibration. Crystals adjusted for a particular frequency are supplied in their holders ; these may be of the fixed gap or variable gap types, and in the case of the latter (Fig. 37), it should be remembered that no alteration or adjustment of any kind can be made without destroying the results of calibration.

When a quartz-controlled transmitter is operating, the quartz crystal increases in temperature, and there is usually a " frequency drift " until the quartz reaches an equilibrium temperature. Moreover, changes in temperature of the space in the neighbourhood of the crystal holder may produce corresponding alterations in frequency. To eliminate these undesirable features, the crystal must be maintained at a constant temperature, except in cases where the temperature coefficient of the crystal is negligibly small—using so-called " zero temperature cut crystals." In some cases temperature control is achieved by enclosing the crystal holder in a small oven, the temperature of which is maintained constant by thermostatic control. Less rigid control is obtained by using the relatively large masses of metal which characterise modern crystal holders (Fig. 37). The large amount of metal acts as a heat reservoir (or sink), and tends to maintain the crystal at a constant temperature, in spite of slight changes in the ambient temperature.

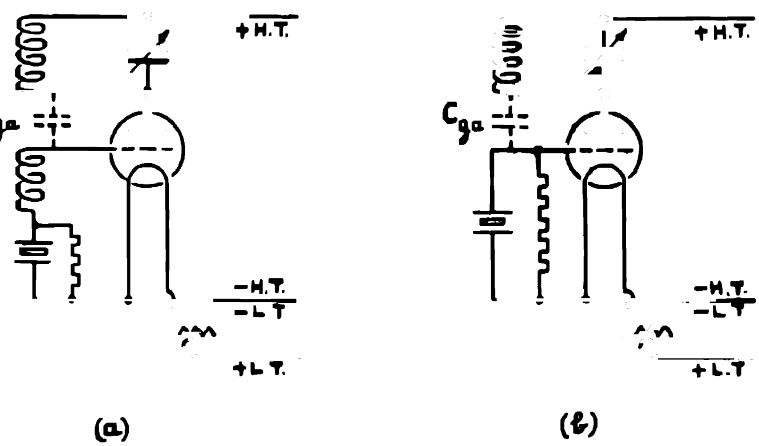
With thermostatic control, the limits of frequency variation may be reduced to several parts in 100,000.

44. Harmonic Picking.—In addition to the fundamental frequency of the crystal, a very long range of "over-tones" may also be generated by suitable circuits. For frequencies above about 6,000 kc./s., it is not always practical for the quartz crystal to be ground thinly enough. In that case,

frequency control can be achieved by choosing the dimensions of the crystal so that a multiple of its fundamental frequency is the one that is required for transmission. That crystal must then be used to control the frequency of a valve working over a non-linear portion of its characteristic ; a long range of overtones will be generated, and the desired one may be selected by suitable circuits and subsequently amplified. Alternatively, resort may be had to " **frequency doubling circuits,**" which are described in greater detail below.

Thus, for a transmission on 12,000 kc /s , a 4,000 kc /s. crystal circuit can be used, since the third harmonic provides the required frequency.

45. **Typical Quartz Control Circuits.**—Two methods of applying quartz control to an oscillatory circuit are shown in Fig. 38 (*a*) and (*b*). The circuit of Fig. 38 (*a*) is of the type described as tuned circuit between anode and filament, mutual inductive grid excitation and series feed, but, if the crystal were not present, the grid excitation would not be sufficient to generate self-oscillations. If, however, a quartz crystal is inserted between the grid coil and filament, the initial oscillatory impulse sets it in oscillation, and the piezo-electric P.D. produced gives sufficient grid excitation to maintain continuous oscillations.



QUARTZ CONTROLLED CIRCUITS.

FIG. 38.

As was suggested in paragraph 42, a vibrating crystal and holder may be considered equivalent to a *highly selective LC circuit* ; it may effectively be replaced by an inductance of very high value in series with a condenser and resistance of very small values, the whole being shunted by a small condenser. It is equal to a stiff electrical circuit having a very sharp resonance peak. Regarded in this way, the resonant frequency of the crystal is obviously slightly altered when other circuits are coupled to the crystal. For example, in Fig. 38 (*a*), the tuned anode circuit acting through C_{ga} will alter the resonant value by an amount depending on the coupling. When the anode circuit is approaching resonance, the small amount of energy fed to the crystal through C_{ga} will be sufficient to start oscillations. These oscillations will increase in value as the tuned circuit approaches resonance, since more energy is fed to the crystal through C_{ga} thus forcing it to oscillate more vigorously but slightly changing the **natural resonance value** of the crystal. If the anode circuit is allowed to approach the resonant state still further, a point will be reached at which the oscillations will suddenly cease. The anode circuit is in parallel with the crystal and constitutes a load which gradually increases the damping of the crystal circuit up to the point at which the positive resistance

exceeds the negative resistance and oscillations then cease. Moreover, it is difficult to maintain the requisite phasing conditions (*cf.* paragraph 16).

The above argument has made it clear that a crystal can only be forced to vibrate vigorously at frequencies very slightly different from its natural frequency. It is, for this reason, more strictly correct to refer to the **over-tones** produced by a crystal oscillator than to employ the term harmonics, which implies that the frequencies in use are exact multiples of the fundamental or natural frequency of the crystal.

When the circuit of Fig 38 (*a*) is to be used over a wide range of frequencies, using different crystals, it may be impossible to avoid sufficient mutual coupling at some frequencies to generate self-oscillations, without the interposition of a crystal. In these circumstances, the crystal no longer controls the frequency, and the circuit of Fig. 38 (*b*) must be adopted. The coupling between anode and grid circuits is then only that due to C_{ga} and there is nothing in the grid circuit except the quartz crystal to affect the frequency.

Fig. 38 (*b*) is the standard circuit in many high-class transmitters. In most cases, a buffer stage of R/F amplification is inserted between the crystal stage and the main R/F amplifier.

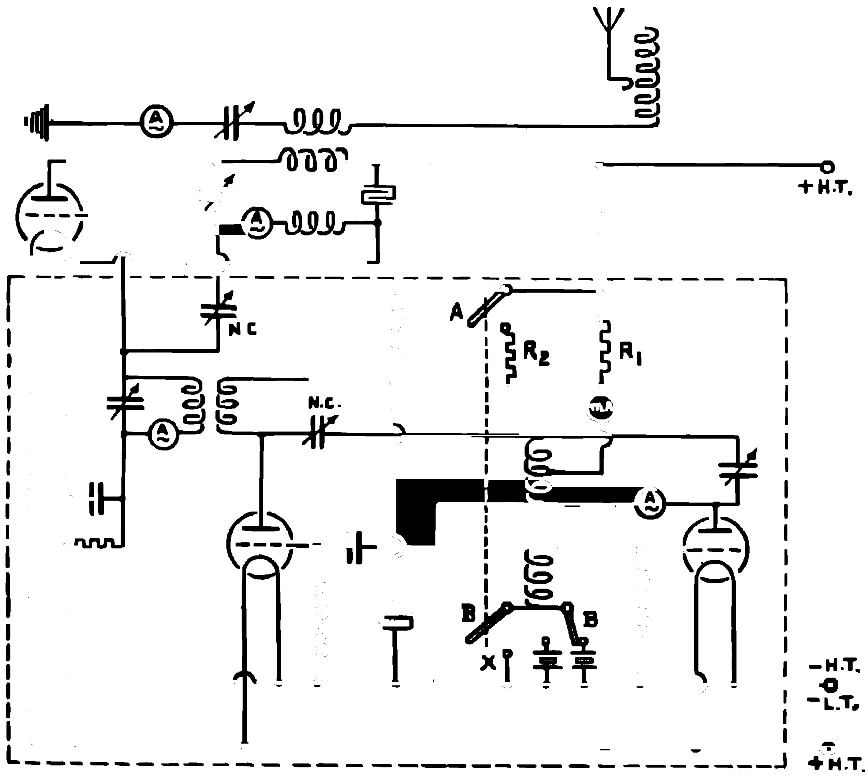


FIG. 39.

The amount of power that can be developed in a quartz-controlled oscillator is limited to a few watts, and when higher power is required the quartz-controlled circuit must be coupled to a power stage, through a buffer stage, as described above for the case of the ordinary master oscillator.

A circuit of this type is shown in Fig. 39, consisting of a master stage, one amplifying stage and a power stage*. Two crystals will be observed, corresponding to two frequencies of transmission. In addition, the master circuit may be converted to a rigid low-power oscillatory circuit at frequencies

* For the proper functioning of the circuits in Figs. 39 and 52, a large R/I by-pass condenser should be joined between H.T. — and the junction of the milliammeter with R_1.

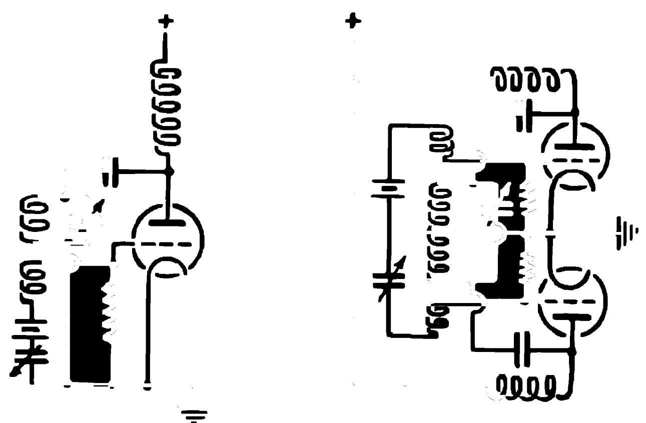
intermediate between those of the crystals, by making switch B to X; by mechanical coupling of the switches this, at the same time, makes switch A, which is open in the other two positions of B.

It has already been seen that the slope of the valve mutual characteristics increases with the anode voltage (Section " B "). With the resistance R_1 alone in the H.T. lead, the steady anode voltage is too low to allow the self-oscillatory condition $\frac{Mg_m'}{C} > R$ to be fulfilled with the values of M (grid coupling) and C obtaining, and the extra grid-filament P.D. provided by the quartz resonator is necessary for oscillations to be maintained. When switch A is made, however, a much smaller resistance, R_2, is put in parallel with R_1, and the consequent increase in steady anode voltage renders g_m' large enough for self-oscillations to take place without the necessity for a crystal.

Half of the oscillatory P.D. across the master tuned circuit is applied between grid and filament of the amplifying valve, the voltage output of which is, in turn, applied by means of the tuned secondary transformer coupling to the grid and filament of the power stage. The output circuit of the power stage is a tuned circuit arranged like the customary Service " divided circuit," with series feed, and direct inductive grid excitation; it is not, of course, self-oscillatory in this case. The power stage is neutralised by the tapped output method (Section " F "); the balancing of the amplifying stage is done by the tapped input method (Section " F ").

In a master controlled circuit, tight mutual couplings may be used without fear of the possibility of frequency jump. It is therefore possible to transfer much greater power to the aerial, and no frequency jump can occur unless the quartz crystal, or master circuit, loses control and the whole circuit becomes self oscillatory.

The quartz control circuit of Fig. 38 (*a*) may be modified for parallel feed as in Fig. 40 (*a*), and adapted to push-pull working as in Fig. 40 (*b*). This type of circuit necessitates the use of two



(a) (b)

FIG. 40.

valves per stage, but has the advantage of being very stable and easily neutralised when working on high frequencies. In a particular case, Fig. 40 (*b*) represents the quartz control stage which is followed by a stage of harmonic picking, and subsequently by power amplification. Fig. 41 shows the circuit details of the first two of these stages. The frequency generated in circuit A is determined by the crystal. The output is applied through the two coupling condensers CC to the grids of the valves in the following harmonic picking stage. The amplified output is passed on to the tuned circuit B which selects the third harmonic of the crystal. Neutralising condensers NN are adjusted so that there is no tendency for energy to feed back from the second stage to the preceding one. As

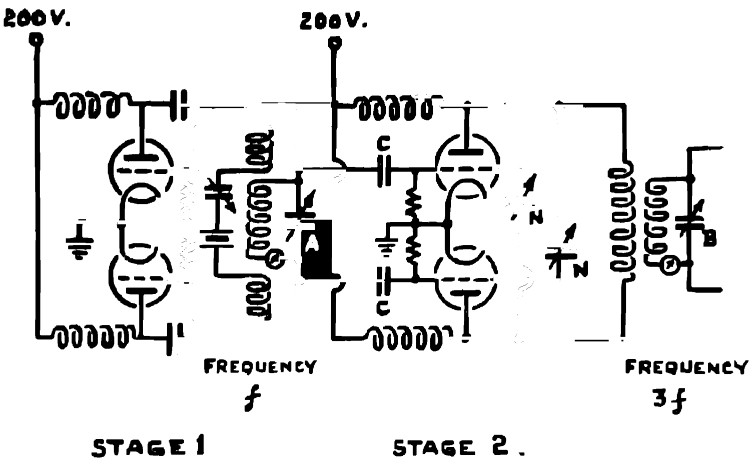


FIG. 41.

has already been observed, in practice the converse is true—that a circuit A when neutralised will be unable to FEED ENERGY FORWARD to a succeeding circuit B, when B has no H.T. applied to its anode. In practice, the latter condition of no feed forward via the inter-electrode capacity is easier to achieve, and the actual process of neutralising consists in adjusting the neutralising condensers until there is no current reading on the ammeter of circuit B when the H.T. supply to stage 2 is broken.

Following the harmonic picking stage are various stages of power amplification.

In practice, an alternative method of neutralising stage 2 is sometimes adopted. The crystal is short-circuited and the quartz stage is rendered self-oscillatory. This is done whenever the " passed on current " from the first circuit, when energised by the crystal, is insufficient to give a reading on the ammeter in the circuit B, rendering it impossible to perform the neutralising process in the ordinary way. Circuit B is tuned until the ammeter shows the maximum reading, and then this current is reduced by means of the neutralising condensers, keeping their values equal. Circuit B is then re-tuned and the neutralising condensers re-adjusted. Having neutralised the two stages for one frequency, they should remain neutralised for an adjacent frequency. The crystal can then be put into operation in circuit A, and circuits A and B tuned for maximum current.

With the advent of H/F pentodes, the process of inter-stage neutralising was considerably simplified. In many cases, the low value of the inter-electrode capacity has made it possible to eliminate the use of neutralising condensers.

46. **Frequency Multiplication Circuits.**—The necessity for frequency multiplication has already been explained in connection with quartz control (paragraph 44). In the push-pull circuit of Fig. 41 the third harmonic was selected, although it is clear that any other higher odd harmonic could have been chosen. It will be recalled that push-pull circuits usually attenuate the even harmonics.

Frequency multiplication may be equally well achieved using one valve only. By working with the bias point at the lower bend, or even beyond the cut off point, the anode current wave form becomes greatly distorted, giving the familiar flattened portions corresponding to the negative half cycles. Such a wave form is rich in harmonics, and it is only necessary to provide the requisite tuned circuits in order to select any particular one.

" Frequency doubling " is only a particular case of the general process, and there are a number of circuits by which it may be achieved. One simple one is shown in Fig. 42 (*a*). The push-pull valves are biased down to the neighbourhood of the lower bend, and it will be assumed that the

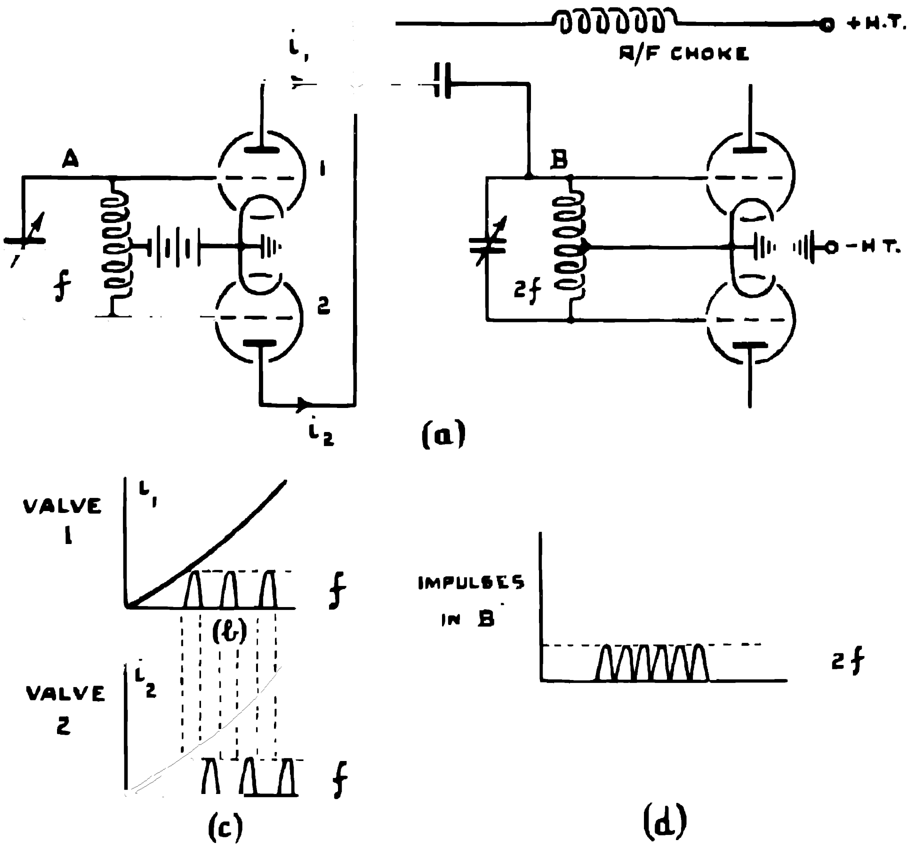


FIG. 42.

circuit A is oscillating at frequency "f" The two anodes have a common H.T. feed, and circuit B constitutes the output impedance between anode and filament, the latter being earthed. Anode current flows during the positive half cycles in each valve; since both anodes are connected together, the pulses of current energise the anode circuit B IN THE SAME DIRECTION. Since two pulses of current flow in each cycle, Fig. 42 (*b*) and (*c*), the output circuit is energised at double the frequency of the input circuit, as shown in Fig. 42 (*d*). If circuit B is tuned to a frequency $2f$, maximum oscillatory current will be obtained. This stage of frequency doubling may be followed by others, or by power amplification.

47. **Special Frequency Stabilised Circuits.**—When it is not desirable to use any form of master control, some degree of frequency stabilisation can be achieved by means of careful design and by the use of special circuits (paragraph 38). In all cases the object is to eliminate or minimise the effect of one or other of the various variable factors in a self-oscillatory circuit. It is proposed to illustrate this by reference to two particular cases, which are not "special" in the sense defined, but which show points of interest.

M/F CIRCUIT.—One of the most essential requirements, is that the oscillatory circuit should be affected as little as possible by variations of aerial or feeder capacity. Any transmitter employing direct aerial excitation can, however, never be free from frequency variation due to the above cause; in particular, the circuit of Fig. 16 (*a*), in which the condenser of the LC circuit may be assumed to be the aerial capacity itself, would need extensive alteration if greater frequency stability were desired. It is essential to use a coupled aerial, and the coupling must be loose, in order that the

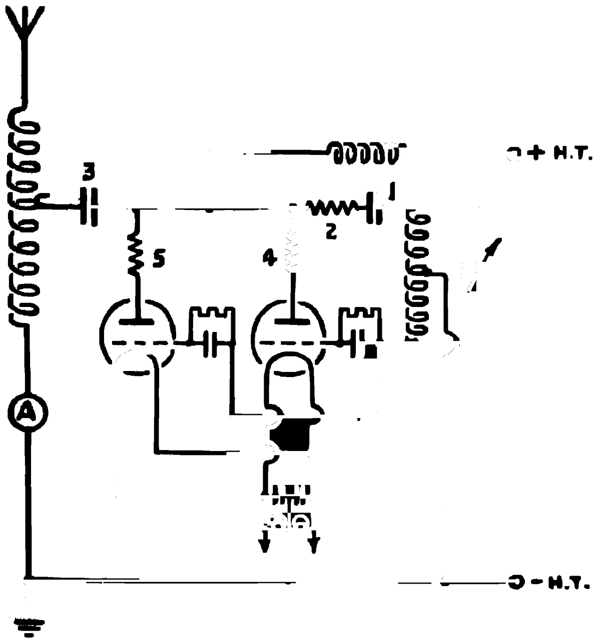


FIG. 43.

tuned circuit may be sufficiently "remote" from the variable capacities in the anode circuit, that the latter may not exercise a controlling influence on the frequency generated.

Fig. 43 represents one way in which the circuit of Fig. 16 (*a*) could be "modernised"; in this case two valves in parallel are used. On the right is seen the self-oscillatory circuit which is of the type classified as tuned circuit between anode and grid, direct inductive grid excitation, parallel feed. It will be observed that the aerial is loosely coupled to the oscillatory circuit by means of the anode blocking condenser (1), the resistance (2) and the blocking condenser (3)

The functions of the resistance (2) and blocking condenser (1) are :—

(*a*) To reduce the coupling between the aerial and oscillatory circuits to a value below that at which the frequency of the complete circuit is determined mainly by the aerial LC circuit; and

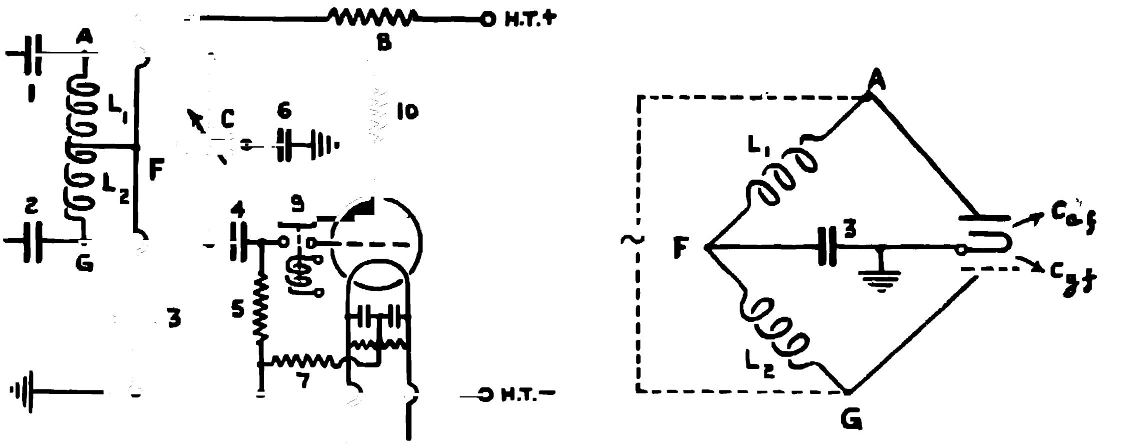
(*b*) To cut down the power supplied to the oscillatory circuit to a value suitable to give sufficient but not over excitation of the grids.

High resistances (4) and (5) are present to minimise the effect of variations in the r_a of the valves; if the added resistances are relatively high, the total effective resistance will remain substantially constant.

Attention must be paid to the design of the various circuit components, and to all of the precautions enumerated in paragraph 39, in connection with the design of master oscillators; under good conditions it may be possible, in this way, to achieve a frequency constancy of the order of $\pm 0 \cdot 1$ per cent.

H/F HARTLEY CIRCUIT.—At H/F the difficulties of designing circuits with any measure of frequency stability, are relatively greater. In general, the circuit must be designed with all the care that is given to a master oscillator; in addition, it is advantageous to provide a circuit which is approximately symmetrical electrically about a central earthy point. Fig. 44 (*a*) represents a circuit essentially simple in appearance, but with certain added refinements. It is a Hartley circuit of the type already discussed, and condensers (1) and (2) represent the capacitive coupling to the aerial (paragraph 28). It is essential, as usual, that the aerial coupling factor should be small. Resistance (8) serves to control the power in the oscillatory circuit; it has another function in connection with the prevention of "blocking" (paragraph 11) which sometimes occurs during the process of keying. Further reference to this feature and to the use of components (9) and (7) of this circuit, will be made below.

In order to achieve a balance about a central earthy point, it is advantageous to use a Kelvin or quadrant type of primary tuning condenser (C), the fixed vanes being connected to the opposite ends of the oscillatory circuit, and the moving vanes being at a floating earthy potential In some cases, in these circuits, the moving vanes are connected to earth through a small condenser (6). The size of this condenser may be arranged so that it is large enough to by-pass to earth some of the



(a) (b)

FIG. 44.

harmonics of the generated frequency. The presence of harmonics is an objectionable feature, for it sometimes happens that the frequency of one of them corresponds to the natural frequency of some closed circuit formed by connecting leads and self-capacities. If this occurs, there will be a drain on energy from the main circuit resulting in a considerable loss of efficiency.

A factor affecting stability, is the position of the anode tapping point F. Mathematically, it can be shown that there is one position of it which considerably reduces the influence of the valve-"constants" in controlling the frequency generated. With that critical adjustment, any change of anode and filament voltage would then have little effect on the generated frequency. This condition remains effective so long as the anode/filament, and grid/filament inter-electrode capacities can be neglected. At the higher frequencies, however, they become of importance and introduce another effect. The valve and the tuning coil of Fig. 44 (*a*) may be re-drawn as a bridge circuit, as in Fig. 44 (*b*). The inter-electrode capacities form two arms of the bridge, and the inductance portions form the other two arms. The bridge will be "balanced" when the ratio of the inductive reactances is equal to the ratio of the capacitive reactances, *i.e.*,

$$\frac{L_1}{L_2} = \frac{C_{gf}}{C_{af}}.$$

If the bridge is not balanced, H/F current will flow through condenser (3) to the filament, and thus vary the emission current. Reducing the value of the blocking condenser (3) will reduce the "unbalance current," if the tapping point F is not correctly chosen, but the point F will then oscillate above and below earth potential, owing to the voltage drop across (3). This affects the correct voltage phase relationships in the circuit, and may even produce "blocking" (paragraph 11).

If blocking occurs, for any reason whatever, the anode current tends to increase to a dangerously high value. This is countered by components (8) and (7); when the anode current increases unduly, there is an increased voltage drop across the resistance (8) and the P.D. applied to the anode is reduced. Moreover, the return space current flows through resistance (7) before reaching the filament; any sudden increase in this current makes the grid effectively more negative with respect to the filament, and so tends to reduce the anode current.

★48. MATHEMATICAL CONSIDERATION OF THE EFFECT OF r_a ON THE FREQUENCY GENERATED.—**In order to make clear that the resonant frequency generated by a valve oscillator does not depend only on the effective LC value of the tuned circuit, it is necessary to treat each particular case in a more exact way than that followed in paragraph 19, and make further deductions during the course of the work.**

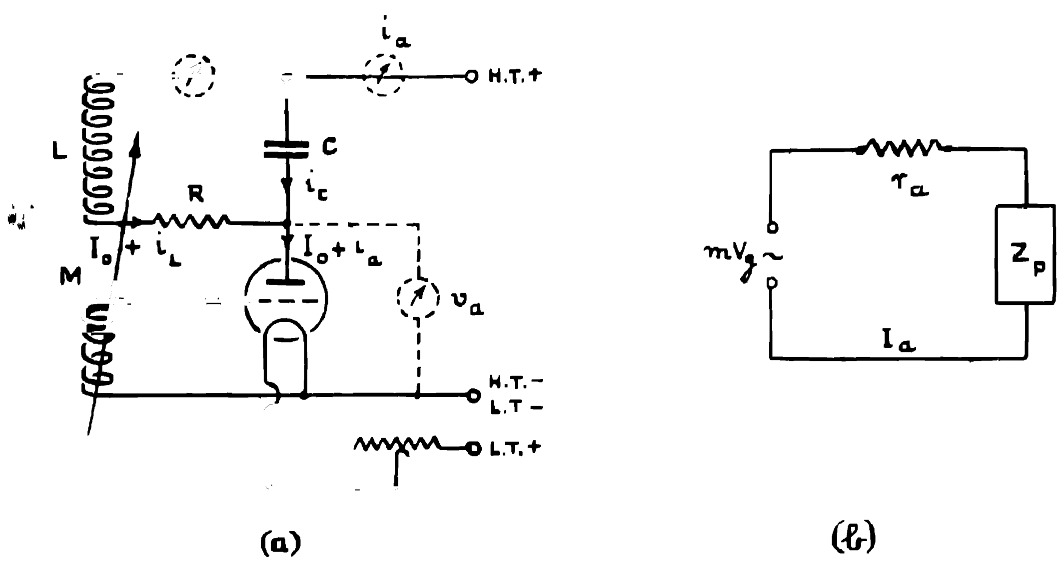


FIG. 45.

Consider the special case of Fig. 3, here reproduced, together with its simplified circuit, in Fig. 45 (*a*) and (*b*). As is shown in Section "F," since the oscillator may be regarded as an amplifier providing its own input, we have, using the usual symbols—

$$mV_g = I_a (r_a + Z_P), \quad \ldots\ldots \quad (1)$$

where Z_P = impedance of the anode circuit.

Now the anode impedance is the sum of two in parallel, Z_C and L_L for the capacitive and inductive arms respectively.

Hence

$$V_a = I_L Z_L = I_a Z_P = I_a\left(\frac{Z_C Z_L}{Z_C + Z_L}\right)$$

$$\therefore \quad I_L = I_a\left(\frac{Z_C}{Z_C + Z_L}\right) = \frac{jI_a}{\omega C\,(Z_C + Z_L)} \quad \ldots\ldots \quad (2)$$

$$\text{since } Z_C = -\frac{j}{\omega C}$$

[Formula (2) may be compared with a similar one which appears in the simple case of two resistances in parallel

$$\text{Also } V_g = j\omega M I_L = \frac{MI_a}{C\,(Z_C + Z_L)} \quad \ldots\ldots \quad (3)$$

From (1) and (3)

$$\frac{MI_a}{C\,(Z_C + Z_L)} = \frac{I_a}{m}(r_a + Z_P)$$

$$r_a + Z_P - m\left(\frac{M}{C\,(Z_C + Z_L)}\right) = 0 \quad \ldots\ldots \quad (4)$$

putting $Z_P = \frac{L}{CR}$ and $Z_C + Z_L = R$, this equation may be solved for M, the critical value of coupling for tl maintenance of oscillations: it is usually called the MAINTENANCE EQUATION, and in its simplest form identical with that found in paragraph 19, namely $M = CR/g'_m$.

More exactly, we substitute in (4) for Z_P, Z_C, Z_L, and proceed as follows—

$$Z_P = \frac{Z_C Z_L}{Z_C + Z_L} = \frac{\left(-\frac{j}{\omega C}\right)(R + j\omega L)}{R + j\left(\omega L - \frac{1}{\omega C}\right)}$$

$$r_a + \frac{\left(-\frac{j}{\omega C}\right)(R + j\omega L)}{R + j\left(\omega L - \frac{1}{\omega C}\right)} - m\left(\frac{M}{C\left[R + j\left(\omega L - \frac{1}{\omega C}\right)\right]}\right) = 0$$

$$r_a R + j r_a\left(\omega L - \frac{1}{\omega C}\right) - \frac{jR}{\omega C} + \frac{L}{C} - \frac{mM}{C} = 0 \quad \ldots\ldots\ldots\ldots\ldots \quad (5)$$

Equating the coefficient of j to zero—the resonant condition—we have,

$$r_a \omega L - \frac{r_a}{\omega C} - \frac{R}{\omega C} = 0,$$

$$\omega^2 = \frac{1}{LC}\left(1 + \frac{R}{r_a}\right) \quad \ldots\ldots\ldots\ldots\ldots \quad (6)$$

This is sometimes called the TIMING EQUATION, and emphasises the influence of r_a on the frequency enerated.

From the real parts of equation (5) we get, the MAINTENANCE EQUATION—

$$r_a R + \frac{L}{C} - \frac{mM}{C} = 0,$$

nd this simplifies to $M = CR/g'_m$, as before.

In a similar manner, the behaviour of any particular oscillator may be analysed, usually starting with a step imilar to that of equation (1)

49. **Inadequacy of Deductions from a Linear Valve Characteristic; non-linear Theory.**—It is possible, not not profitable, to follow up the deductions of the last paragraph. Vith reference to equation (6), one can devise special circuits to eliminate the effect of R and r_a, nd render the frequency substantially dependent only upon the LC value. These circuits usually ork well for one particular frequency, but troubles due to frequency variation are often exaggerated. n practice, it is quite futile to attempt to produce a rigidly controlled oscillator by making use of ircuits dependent on the results of deductions based on the assumption of a linear valve character- tic.

For approximate work, a linear treatment of the valve oscillator yields simple and useful esults; all of the work in this section is based upon linear considerations using—

$$I_a = g_m V_g + \frac{V_a}{r_a} \quad \ldots\ldots\ldots\ldots\ldots \; cf.\ \text{K.19.}$$

Vhere requirements demand something better than this, it is essential to employ non-linear valve neory I.C.W. wavelts associated the n rsome mathematics. The insufficiency of a linear theory of istainers. An audio frequency beat n Rayleigh in 1883. Some of his work has received consider- ble extensio modulating frequency " by the hands of B. van der Pol, Appleton, and others; after omewhat involve correspond to the me ing, it can be shown that a valve oscillator can give a his is cons oscillation Section " N ." iable form, and that the pure sine wave is a special limiting one of the ole results of analysis. Linear theory can only account for The produc ns which e A2 waves, ne waves in shape; non-linear theory can give a precise xplanation It shou odic oscid that which is not sinusoidal, and account for the conditions which mit the amplitu W. oscillations (K.20). Moreover, a similar treatment provides an explanation the cases which o For a succession of non-periodic phenomena, called " relaxation oscillations " y van der Pol. A3 w

In practice, the effect of the curvature in the valve characteristic is an apparent increase in the power factor of the circuit. Instead of the anode oscillatory volts V_a, at the fundamental frequency, being 180° out of phase with I_a (*i.e.*, V_R being in phase with the anode oscillatory current, *cf.* K.6), it differs by a small amount depending on the curvature of the valve characteristic, and on the nature of the external circuit. A further consideration of this subject is outside the scope of this work.

50. Self Oscillatory Circuit with Capacitive Feed Back.—Fig. 46 (*a*) represents a self-oscillatory circuit of a type which has not previously been considered, and which has certain points of interest.

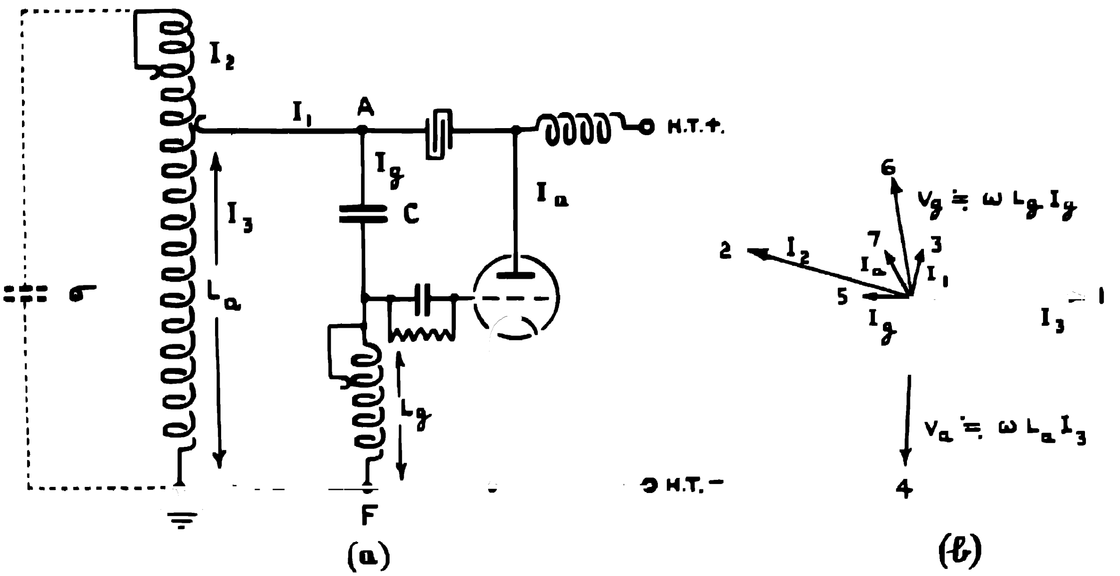


FIG. 46.

The tuned circuit consists of the aerial inductance and aerial capacity, but the circuit oscillates as a whole, the LC value of the aerial circuit being influenced by the condenser C and grid exciting inductance L_g. The magnitude of the " coupled in " capacity and inductance, depends upon the transformer ratio of the auto transformer coupling, a subject which is fully treated in the section concerned with simple transformer theory. In standard terms, the description of the oscillator would be—tuned circuit between anode and filament, parallel feed, capacitive reaction (or feed back) and direct inductive grid excitation, direct inductive aerial excitation. The grid exciting inductance is made variable, and, under operating conditions, it may be found that the nett reactance of the arm AF is always capacitive.

A more precise explanation of the reason why this circ[illegible] with the help of the vector diagram of Fig. 46 (*b*). Quite clearly, if [illegible]tive reactance, with reference to any oscillatory voltage developed b[illegible] oscillatory voltage developed across the grid exciting inductance will be 180 [illegible] V_a and V_g approximately 180° out of phase, thus establishing the fu[illegible]dition for [illegible] The complete vector diagram may be built up by consid[illegible]s in the [illegible] they are numbered in the diagram. Commencing with vector I_3, the condenser cur[illegible] shown leading it by an angle which is just less than 180°, in order to allow for the effects [illegible] value [illegible]tance. The vector sum of these two currents is I_1, the supply current to the rejector c[illegible] Between the anode tapping point and earth, the current I_3 develops a back E.M.F. i[illegible] inductance L_a. This

back E.M.F. may be regarded as being responsible for the cu... oscillatory valve anode by a valve-
being regarded as a circuit connected in parallel with a portion
back E.M.F. is V_a, and in the diagram it is drawn slightly more tha...
...clude the effects of resistance. The current I_g may then be drawn lagging ...requency oscillation is main-
V_a. From this we can draw the position of the vector V_g lagging approxima... of the type classified as
... will therefore be almost in anti-phase to V_a. The vector sum of currents I_1 and ... feed, and anode at
...y vector 7, which represents the anode oscillatory current I_a. It therefore appears tha... taken directly
...re almost in phase, and the system is thus a self-supporting one, capable of producing ...ndenser C_1
...scillations. It should be noted that the maintenance of this condition depends upon the pre... C_1 has a
...tion of nett capacitive reactance in the arm AF.

51. **Modulation; Types of Waves.**—All of the transmitters or oscillators so far described in this section have been designed to produce a radio frequency current of constant amplitude in the aerial circuit, except in so far as that current may be interrupted by keying for signalling purposes. The type of wave produced by such transmitters is called C.W.—continuous wave— ; outside the Service it is sometimes known as type A1, and defined as follows :—

TYPE A1.—Continuous waves, unmodulated, key controlled. Continuous waves of which the amplitude and/or frequency is varied by the operation of keying in telegraphic transmission.

The keying of valve transmitters will be described later ; it reduces the amplitude from a constant value to zero, *i.e.*, it gives amplitude variation. An instance of frequency modulation was seen in K.35. The reception of Type A1 waves necessitates the use of a local oscillator, *i.e.*, heterodyne reception, at the receiving station, or of some equivalent method of rendering the signal perceptible to the senses.

An alternative method of signalling is to impress an audio-frequency variation (or, in certain cases a supersonic frequency variation) on the frequency of the radio frequency waveform as radiated from the transmitter. An example of this has already been seen in spark transmission, in which case the spark wave trains follow each other at an audio frequency. The damped waveform radiated from a spark transmitter falls under the class known as TYPE B waves, and defined as waves forming successive wave trains, in each of which the amplitude after reaching a maximum progressively decreases.

The other types of waves of this nature are classified as Type A2 and Type A3, and defined as follows :—

TYPE A2.—Continuous waves in which a variation of amplitude and/or frequency is made in a periodic manner at an audible frequency (or in certain cases at a supersonic frequency), and key-controlled for the purposes of telegraphic communication.

TYPE A3.—Continuous waves in which a variation of amplitude and/or frequency is made in accordance with the characteristic vibrations of speech or music.

These different waveforms are shown in Fig. 47. In the Service, Type A2 waves are usually called Interrupted Continuous Waves (I.C.W.). In the particular case where the variation of amplitude or " modulation " is sinoidal, the kind of I.C.W. produced is called " Tonic train." The use of I.C.W. waveforms in which the modulation is supersonic, is described in the section on Quench Receivers. An audio frequency beat note is obtained if the " quench frequency " differs from the supersonic " modulating frequency " by an audible amount.

Type A3 waves correspond to the method of communication known as Radio Telephony (R/T) ; this is considered in Section " N ."

The production of Type A2 waves, or I.C.W., by valve transmitters, will now be described in some detail. It should be noted that there is no essential difference between the simple modulations which we call I.C.W., and the more complex modulations corresponding to the various notes of speech or music. For this reason, the same general principles apply both to the production of I.C.W. and to Type A3 waves.

In practice, the effect of the
the power factor of the circ
frequency, being 180° ou
cf. K.6), it differs b
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this work.

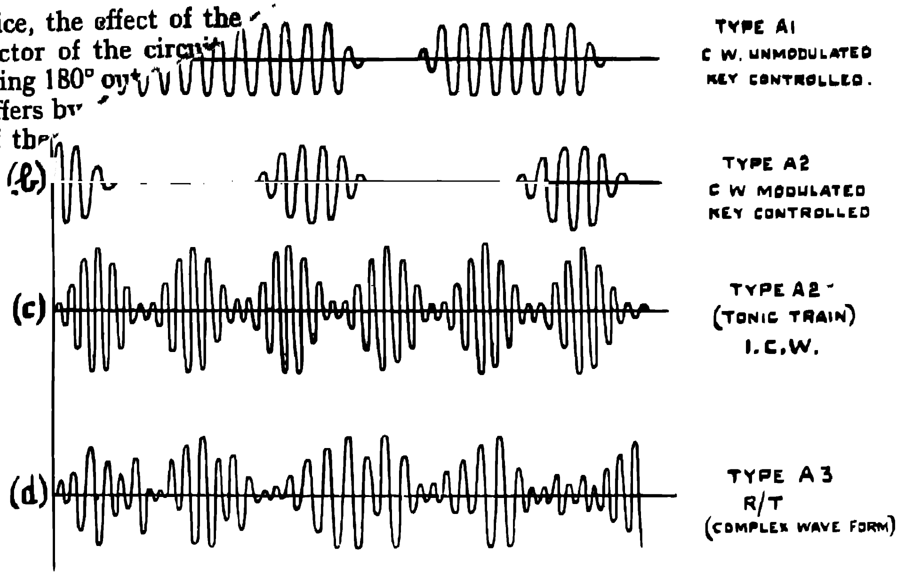

FIG. 47.

52. **Methods of Modulation.**—It has been seen in K.10, that the power developed in the tuned circuit of a valve generator, and hence the amplitude of the oscillatory current, depends on several factors, including the following :—

(*a*) The steady H.T. voltage applied to the anode—anode modulation.
(*b*) The steady potential of the grid—grid modulation.
(*c*) The relation between the impedance of the tuned circuit and the A.C. resistance of the valve—modulation by absorption.

The alteration of any one of these factors produces a change in the amplitude of the oscillatory current. If the alteration has an audio frequency variation whose amplitude corresponds to that of a musical or speech vibration, the R/F oscillation will therefore be modulated in amplitude in a similar manner, and a waveform whose amplitude varies at audio frequency will be radiated.

The methods of modulation in common use depend on the three factors enumerated above, the most usual being ANODE MODULATION. These general methods of modulation apply also to the production of Type A3 waves.

The methods employed in the Service for producing I.C.W. are as follows :—

53. **Methods of Producing I.C.W.**—(1) The transformer secondary voltage from an alternator and transformer arrangement may be applied directly between anode and filament of the transmitting valves—ANODE MODULATION.

A circuit for this purpose is shown in Fig. 48. The transmitter is of the type described as tuned circuit between anode and filament, mutual inductive grid excitation, parallel feed, direct aerial excitation, and the H.T. supply is an alternating voltage of half the amplitude of the transformer secondary voltage.

The amplitude of the radio-frequency oscillations follows roughly the transformer secondary waveform during the positive half-cycle. This is shown as a sine curve in Fig. 49, but the remarks in H.9 should be noted. During the negative half-cycle the anode is negative to the filament and no self-oscillations can be generated in the transmitting circuit.

As there is no smoothing condenser, the transforequency oscillatory valve anode by a valve-
for radio-frequency currents, otherwise large radio-frequency
tion, may be developed across it (the inductive reactance of the radio-frequency oscillation is main-
adio-frequencies). The self-capacity of the winding is usually sufficibeing of the type classified as
by-pass condensers may be shunted across the two halves of the seconaaseries feed, and anode at
'is not taken directly
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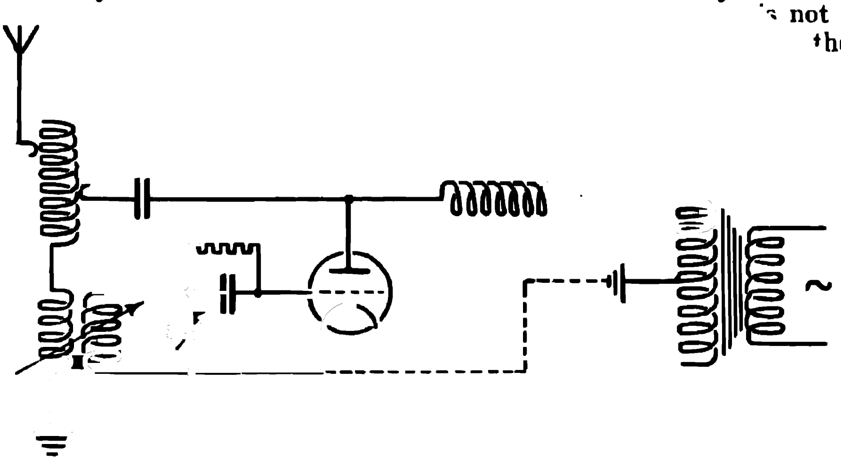
it. C_1 has a



SINGLE PULSE I.C.W. CIRCUIT.

FIG. 48.

fitted, even in C W transmitters, to by-pass any radio-frequency currents that may find their way to the transformer secondary by means of stray coupling to the H.T. leads. The reactance of these condensers is large enough at the **alternator** frequency not to affect the secondary P.D.

As will be seen from Fig. 49, the A/F variation of amplitude is by no means sinoidal and, moreover, it suffers further distortion due to the rectified D.C. component flowing through the transformer secondary winding and partially saturating the core. The audible result in the receiver is more of a noise than a note. The A/F variation occurs at the frequency of the alternator, and the waveform radiated is called **single pulse I.C.W.**

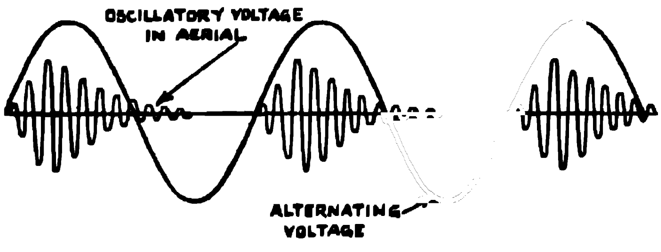


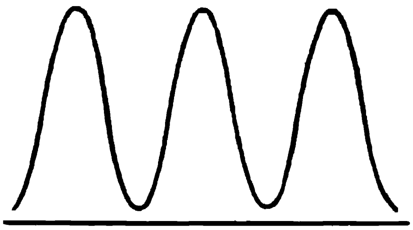
FIG. 49.

Service circuits are commonly designed to transmit either C.W. or I.C.W., so that the transformer secondary is across the anodes of the rectifying valves (Section "H"). The modifications required when changing over from C.W. to I.C.W. are to break the rectifier filament heating circuit and to transfer the H.T. lead from the positive plate of the smoothing condenser to one end of the transformer secondary.

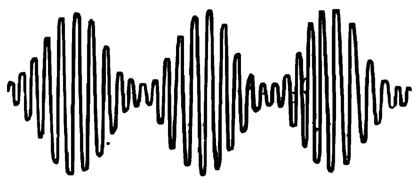
Keying for telegraphic transmission may be arranged by a key in the transformer primary circuit—PRIMARY KEYING.

...itage with the smoothing condenser removed—ANODE

In practice, the effect of the ...
the power factor of the circ... ...at the ripple, or percentage change, in the D.C. voltage across the
frequency, being 180° out ... the transmitting valves was inversely proportional to the capacity
cf. K.6), it differs b... ...en the condenser is cut out altogether, the ripple on the rectified supply
the nature of the ...u, and the voltage applied to the transmitting valve will be of the form illustrated
this work.



VOLTAGE FROM RECTIFIER.



CURRENT IN OSCILLATORY CIRCUIT.

FIG. 50.

The voltage from the rectifier never quite falls to zero, on account of the capacity to earth of the filaments and the inductance of the anode choke, which tend to keep the current flowing during the intervals between the current pulses from one or other of the two rectifying valves.

It should be noticed that, if full-wave rectification is employed, the variations in the supply voltage occur at **twice** the frequency of the A.C. supply, and so this method gives a note whose frequency is twice that of the note given by the previous method.

For this reason the waveform radiated is called DOUBLE PULSE I.C.W.

Keying is effected, as before, in the primary circuit of the transformer.

If half-wave rectification is used, SINGLE PULSE I.C.W. is produced.

(3) Mechanical methods for interrupting the generation of oscillations at a frequency corresponding to audibility.

A simple method of producing I.C.W. by this means is to make and break the grid lead of the oscillating circuit, by inserting a buzzer wheel in series with the grid leak. When the buzzer wheel is bearing on a conducting segment, an oscillation will be set up in the aerial, but when it bears on an insulated segment the grid will be left insulated and will accumulate such a negative charge that the aerial oscillation will be quenched. The result will be I.C.W. of a frequency depending on the speed of revolution of the buzzer wheel. During the time that oscillations are being generated, their amplitude is sensibly constant, and the waveform corresponds to that of Fig. 47 (*b*). I.C.W. produced in this manner is sometimes called "chopped C.W.," a terms which is explained by the waveform. This method is suitable for use with an A.C. source of supply whose frequency is too low to give a good note. A wide frequency spread is produced, and the method is only used when great economy of space is essential.

The three methods hitherto given produce variations in amplitude at audible frequency which are not sinoidal, and cannot be said, therefore, to give the special type of I.C.W. known as "TONIC TRAIN." This latter is only given by the fourth method, now to be examined.

(4) **Modulation of the H.T. supply to the radio-frequency oscillatory valve anode by a valve-maintained audio-frequency voltage.**

A typical circuit for this purpose is shown in Fig. 51. A radio-frequency oscillation is maintained in V_1 and its accompanying tuned circuit, the arrangement being of the type classified as tuned circuit between anode and grid, direct inductive grid excitation, series feed, and anode at high oscillating potential. The H.T. supply to the anode of V_1, however, is not taken directly from a D.C. or rectified A.C. source, but is, in fact, the voltage developed across the condenser C_1 due to an audio-frequency oscillation maintained in the valve V_2 and its tuned circuit. C_1 has a

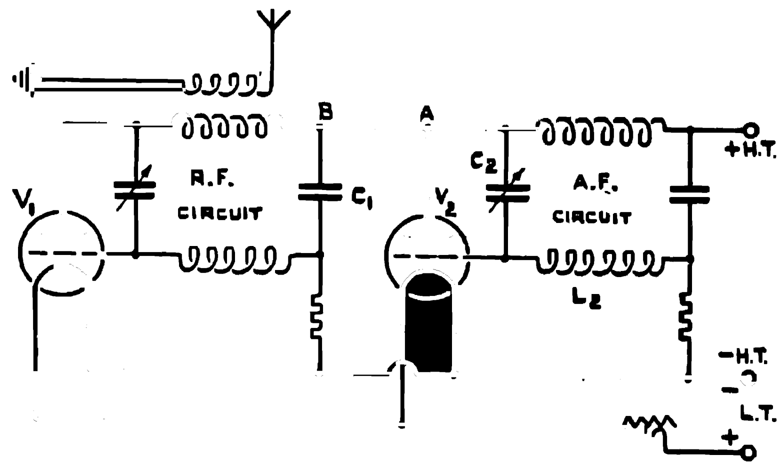


FIG. 51

reactance to audio-frequencies of the same order as C_2, and is in parallel with C_2 and L_2 in series ; it must therefore be considered as part of the audio-frequency tuned circuit, and an audio-frequency P.D., proportional to the oscillatory P.D. between the anode and filament of V_2, is impressed across it. Thus the H.T. supply to the anode of V_1 is modulated at audio-frequency, and the R/F oscillations vary correspondingly in amplitude. Since the P.D. across C_1 can be arranged to be very nearly sinoidal by applying the appropriate grid bias to the grid of V_2, the R/F amplitude modulation is likewise sinoidal, *i.e.*, tonic train is produced. The depth of modulation, *i.e.*, the ratio of the amplitude of the audio-frequency P.D. across C_1 to the amount of the steady H.T. supply, also requires careful adjustment if sinoidal modulation of the R/F oscillations is to be achieved. This point is discussed more fully in Section " N."

To increase the P.D. across C_1, a choke is sometimes added between A and B, as in Fig. 52, of such a value that the path through C_1 from A to filament is partially tuned to resonance at the audio-frequency produced by V_2. A greater audio-frequency current then flows through C_1, and a correspondingly greater P.D. is produced across it, provided that the total oscillatory P.D. from A to filament is not too much decreased by the insertion of the choke. This, however, is liable to be the case since the impedance of the V_2 tuned circuit between anode and filament has thereby been considerably reduced. If tuned to resonance, for example, the choke and C_1 would constitute a path of very low impedance in parallel with the remainder of the V_2 tuned circuit, and there would be a large departure from the most efficient condition, viz., that the said impedance should approximate in value to the A.C. resistance of V_2.

The valve V_2 is also sometimes omitted, either C_1 or C_2 being removed, the free end of the grid inductance of the R/F circuit being connected to the grid inductance of V_2. V_1 then has to maintain both the A/F and the R/F oscillations. It is difficult to obtain a circuit of this type which operates satisfactorily ; even if oscillations can be maintained, the efficiency is poor.

Fig. 52 shows a complete high frequency I.C.W. transmitter, in which the master circuit can either be used as a rigid low power self-oscillatory circuit, or quartz control can be employed; it will be observed that the circuit is the same as that of Fig. 39 with the addition of the A/F modulating circuit.

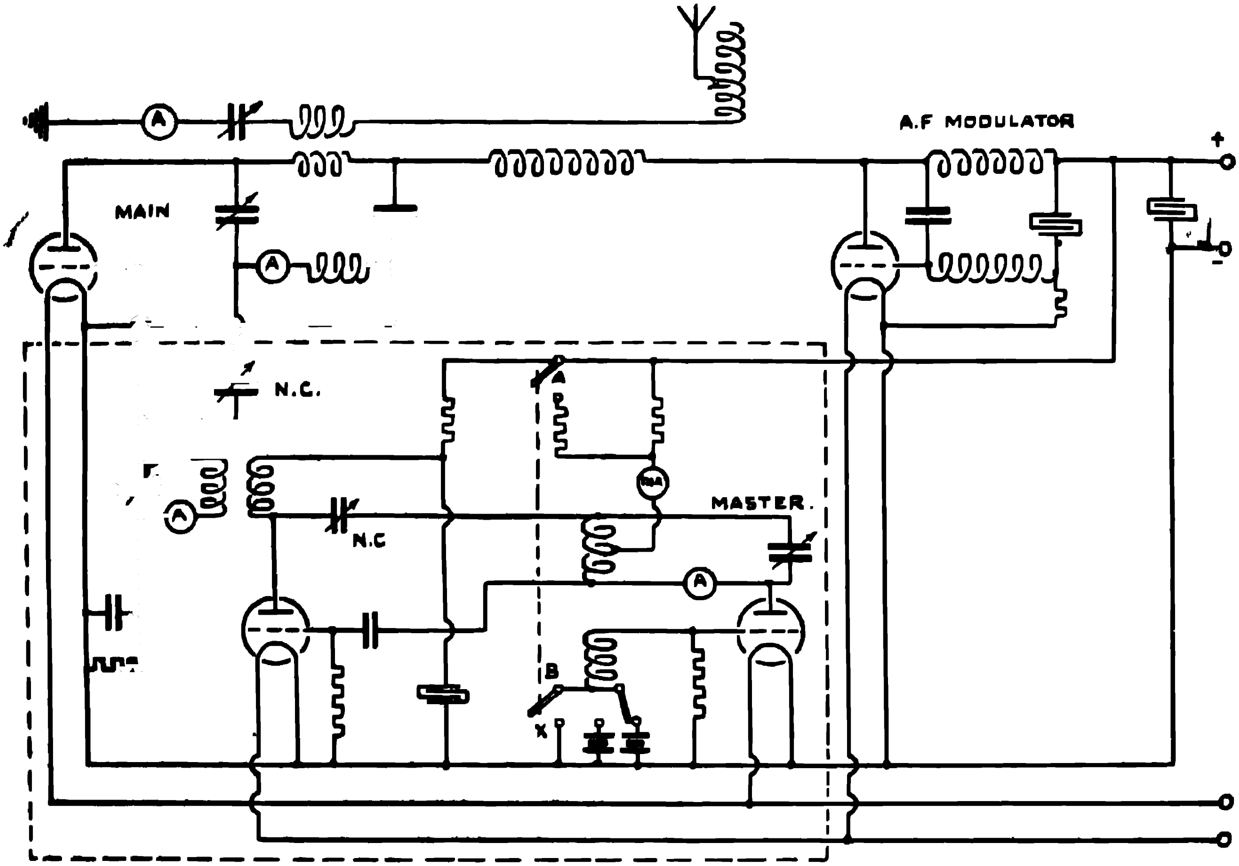


FIG. 52.

(5) The circuit diagram of a Service V.H/F transmitter producing Tonic Train is shown in Fig. 53. The R/F oscillatory circuit is enclosed by the dotted line. The oscillating valves are arranged in push-pull, the tuned circuit being connected as usual between their anodes. With respect to either valve, however, the tuned circuit is between anode and filament, the H.T. supply lead being the filament connection. Series feed is employed, and the anodes are the high oscillating potential electrodes. The aerial coupling is mutual inductive.

There is no inductive coupling between the grid coil and the tuned circuit inductance, the grid excitation being obtained by direct capacitive coupling through the anode-grid inter-electrode capacities of the two valves. It was seen in Section "F" that an amplifier stage with a tuned grid circuit and an output choke was liable to generate self-oscillations; this tendency is, of course, unaffected if the positions of the tuned circuit and the untuned choke are reversed, and at very high frequencies is sufficiently pronounced to give a satisfactory transmitting circuit.

The modulation system employed is that variety of anode voltage modulation known as "CHOKE CONTROL" and described in Section "N." A high impedance 1 : 1 transformer is used instead of a choke. This has the advantage that the direct currents to the oscillator and main

modulator valves can be arranged to flow in opposite directions through the two windings, and so produce steady fluxes in opposite directions in the core. The resultant magnetisation of the core is the difference of these two effects, and so the distortion produced by a large steady magnetisation of the transformer core is avoided.

The A/F oscillations are generated in a circuit of the familiar type classified as tuned circuit between anode and grid, direct inductive grid excitation and series feed. The H.T. supply to the anode is cut down to an appropriate amount by a resistance in the anode lead. The A/F oscillation is amplified by two main modulator valves in parallel, before being applied to the anode of the oscillator valve.

In order to produce tonic train modulation, the same precautions against distortion must be taken as those mentioned in Section "N."

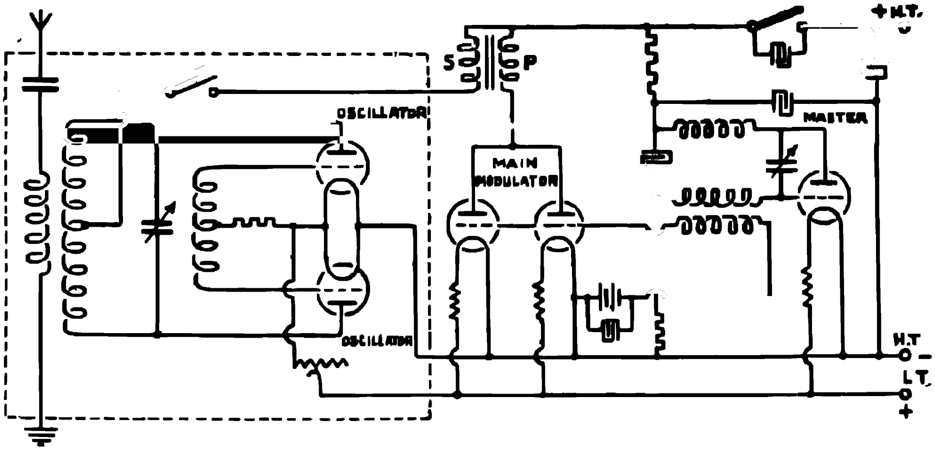


FIG. 53.

(6) With the exception of the mechanical method for the production of I.C.W. described above, instances of grid modulation have not yet been seen.

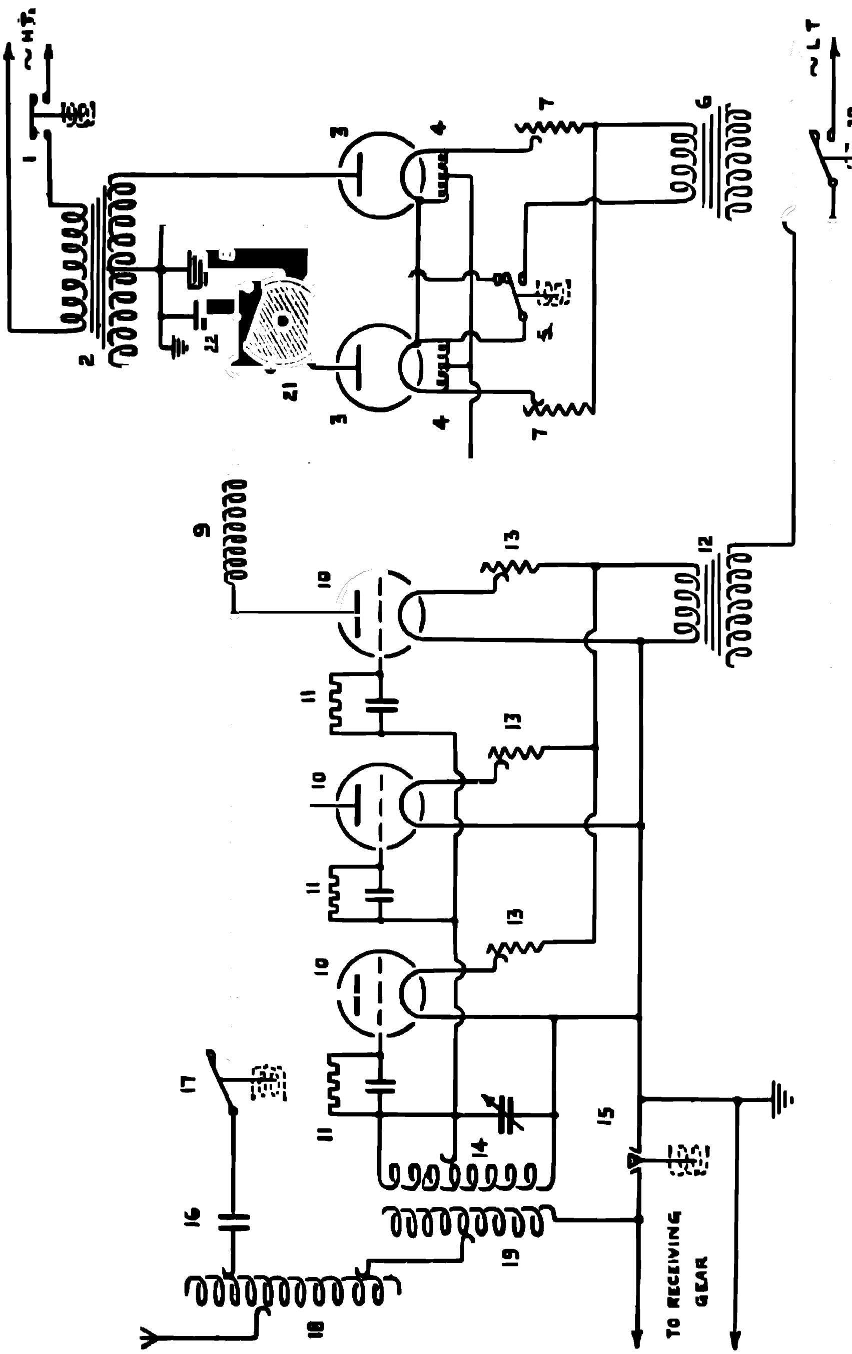
There are several ways of using grid modulation, some of which are particularly applicable to the production of R/T and will be described in the appropriate section. Type A2 waves are produced by a self-quenching oscillator—a squegger; a full account of this action is given in Section "F," and it is only referred to here for the sake of completeness.

54. **Complete Self-Oscillatory Valve Transmitting Circuit.**—In Fig. 54 is illustrated a typical self-oscillatory valve transmitting circuit for producing either C.W. or I.C.W. According to the classification given earlier in this chapter, it is described as :—

Tuned circuit between anode and filament.
Mutual inductive grid excitation, the grid being partially tuned.
Parallel feed.
Direct aerial excitation.
Three valves in parallel, their anodes being the high oscillating potential electrodes.

The H.T. voltage on the anodes is obtained from a rectifier unit employing full wave rectification of an A.C. supply.

When I.C.W. is required, the smoothing condenser (8) is cut out of circuit by means of the C.W.-I.C.W. switch (21).



SELF OSCILLATORY VALVE TRANSMITTING CIRCUIT COMPLETE

FIG. 54.

The components indicated by numbers are as follows :—

(1) The magnetic key to interrupt the A.C. supply for signalling purposes.
(2) The H.T. step-up transformer.
(3) The rectifying valves.
(4) The equaliser coils.
(5) The rectifier switch for completing the rectifying valve filament heating circuit, and for discharging the smoothing condenser (through the equaliser coils) when transmission is finished.
(6) The step-down filament transformer for the rectifying valves.
(7) The rheostats for controlling the filament circuit of the rectifying valves.
(8) The smoothing condenser.
(9) The anode choke.
(10) The transmitting (or oscillating) valves.
(11) The grid insulating condensers and leaks.
(12) The step-down filament transformer for the transmitting valves.
(13) The rheostats for controlling the filament current of the transmitting valves.
(14) The partially-tuned grid excitation circuit. The idle turns of the inductance are short-circuited.
(15) The operating switch (across which the receiving gear is joined), which completes the earth for the aerial circuit during transmission.
(16) The anode blocking condenser.
(17) The anode key, which breaks the circuit between anodes and aerial when the transmitting key is not pressed, thus preventing a path to earth through (18), (16), (9) and (8) for received signals.
(18) The aerial coil.
(19) The aerial coupling coil.
(20) The filament switch for breaking the primary supply to the filament transformers.
(21) The C.W.-I.C.W. switch.
(22) The small condenser to allow the R/F. current through the anode choke to pass directly to earth (or L.T.—) in the I.C.W. position. Since the smoothing condenser is then out of circuit, the easiest path for this current to earth would otherwise be across the windings of the filament transformer, which might therefore be damaged.

55. **Methods of Signalling.**—For telegraphic transmission, the C.W. or I.C.W. oscillations produced by a transmitter must be broken up into wave trains, the duration of which is the same as that of the dots and dashes of the morse code. This implies that the transmitter must be periodically started and stopped, the whole action being controlled by means of a key suitably placed. There are many ways in which this may be done, and the most suitable method for use in a particular case depends upon (*a*) the power of the transmitter, and (*b*) the requisite speed of signalling.

A hand key suitable for signalling, can only be directly inserted in lower power circuits, and the general practice is to make and break some lead in the circuits by means of a magnetic key, which, in turn, is operated by the hand key. Oscillations are maintained while the key is pressed and cease when it is released. This naturally leads to transient conditions at the instant when contact is made or broken by the key, so giving rise to " key clicks " in the received signal and variations in the transmitted frequency.

In Service transmitters, the key may be found in various positions. In high powered sets, supplied by rectified A.C., it is placed in the transformer primary, as in the case of Fig. 54. In sets supplied by D.C. enerators, it is normally in the positive H.T. lead to the anode, as in Fig. 53.

Sometimes the key is inserted in the H.T. lead to the filament, between the filament itself and the point where the grid leak is attached to the filament lead. Breaking the key then breaks the anode supply, and, at the same time, isolates the grid from the filament. The filament continues to shoot out electrons and becomes highly positive with respect to the grid, *i.e.*, the grid is made

relatively highly negative, and oscillations are quickly damped out. For high speed signalling, the grid may be made to go negative at a still faster rate in order to reduce the transient period ; keying should then be confined to the output side of the rectifier unit, so as to avoid the lag which arises when the smoothing condenser has to be charged anew for each wave train.

It is possible to distinguish four fundamental methods of signalling—Primary signalling, High tension signalling, the Marking and Spacing waves method, Grid signalling. These are described in further detail below.

56. **Primary Signalling.**—This consists in making and breaking the supply to the transformer primary circuit, as in Fig. 54. In this method the whole of the power supplied to the set has to be controlled by the key. If the set is one of any appreciable power this involves a well-designed key with considerable mechanical and electrical inertia ; this would be a serious disadvantage if high speed signalling were required. In general, the type of key fitted must be capable of breaking a high current at a comparatively low voltage. An advantage is that no power is wasted when the key is not pressed ; a disadvantage is that the smoothing condenser cannot be large.

57. **High Tension Signalling.**—An alternative method of cutting off the supply to the transmitter is obtained by fitting the key in the high tension line, *i.e.*, in the secondary circuit of the transformer. The power to be controlled is the same as in the first case, but the current is less and the voltage is correspondingly greater. In general, this means that the key must be bigger and it is sometimes necessary to use blowers to prevent arcing across the key contacts. A disadvantage is that the voltage across the rectifying condenser rises during the "space" periods ; an advantage is that the condenser can be large.

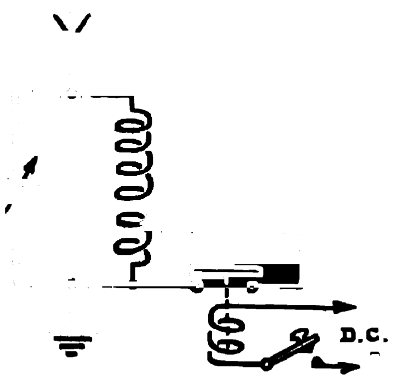


FIG. 55.

58. **Marking and Spacing Waves.**—In this method of signalling, the transmitter is constantly in operation, but the frequency is varied between two frequencies sufficiently far apart, in accordance with the symbols of the morse code. The rapid change of frequency may be achieved by short-circuiting a portion of the tuned circuit inductance by means of a key, as shown in Fig. 55. This method is one which is open to many objections, of which the radiation of two wave frequencies is by no means the least important. This method gives a very slow maximum speed of signalling and has little to recommend it.

59. **Grid Signalling ; High Speed Methods.**—In order to appreciate the fundamental problems of high speed signalling, it is convenient to consider what happens during the radiation of a morse symbol.

When the key is on open circuit, the rectifier and its associated machines are working under no load conditions ; unless the voltage is stabilised (H.14), the rectifier output will be higher than that under steady load conditions. Assuming the latter to be the case, when the key is pressed the rectifier output volts will fall exponentially, and the aerial current will rise in a similar manner, both processes being approximately represented by Fig. 56. The diagram shows that the growth of aerial current is spread over a period of time, the actual amount depending upon the inertia of the whole circuit. The flat part of the two curves represents the state of affairs in the steady state after the initial transient conditions. If the key is broken at C, the aerial current will not immediately fall to zero but will fall exponentially along the slope CD at a rate again depending on the circuit inertia. The duration of the morse symbol is therefore given by the distance AD along the time axis.

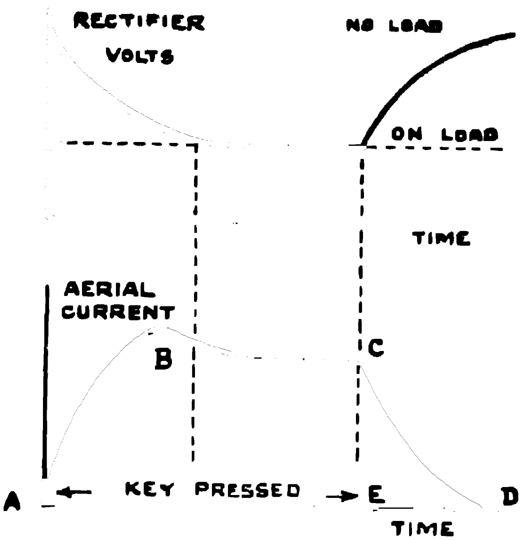


FIG. 56.

As the speed of signalling is increased, the duration of a symbol is correspondingly shortened, and the points C and B approach each other. In the case of a rectifier with no stabilising device, and neglecting the time of operation of the key, the speed of signalling will be the highest possible when the points B and C coincide.

From this it follows that in primary signalling, the speed is limited by the mechanical inertia of the key and the electrical inertia of the circuits. In practice, the maximum reliable speed which can be obtained with a normal design of primary key, corresponds approximately to the maximum speed of manual operation, *i.e.*, 30-35 words per minute.

To achieve speeds higher than the above, it is necessary to minimise the inertia of the key and to eliminate as far as possible the electrical inertia of the circuit. These two conditions can be achieved by the method of signalling which has become known as GRID SIGNALLING. In principle, it consists in—

(*a*) placing the key in the grid circuit, so that it directly controls very little power and can therefore be small, and

(*b*) partially eliminating the transient conditions at the beginning and the end of the morse symbol by means of suitable voltage stabilisation.

Signalling by means of a key in the grid circuit depends upon the delicate control of the anode current of an oscillating valve which is obtained by varying the grid potential. The signal is divided into two periods. When the key is pressed, the valve oscillates and we have the MARK period; when the key is released a heavy negative bias is applied to the oscillating valve, biasing it to a point beyond the cut-off, so that the valve ceases to oscillate, and we have the SPACE period.

Voltage stabilisation is achieved by arranging that the load on the rectifier is the same during the mark and space periods. During MARK, the oscillating circuit constitutes the load; during SPACE, the ABSORBER VALVE and associated anode resistance provide an equivalent load. Power absorption during the space period is also necessary with high speed keyed to avoid the possibility of power line surges, due to electrical resonance of the supply circuits. Furthermore, it reduces "key clicks" and frequency variation. Its only disadvantages are the extra cost of valve rectifiers, and the power waste during *space*.

Fig. 57 represents diagrammatically a low speed grid signalling and absorber valve arrangement used in the Service. Valve (1) will be assumed to be the master control stage of a transmitter, and valve (2) is the absorber valve, the latter being capable of taking a load equal to that taken by the former. It should be noted that, in the interests of frequency constancy, it is recognised to be bad practice to key the master stage. In Service low speed signalling it is, however, necessary in order to make it possible to "listen through"; for an operator to receive with ease the necessary procedure signals during a transmission, it is essential that interference from sources such as oscillators should be minimised. When the latter is not a consideration of importance, as in high speed signalling, it would be better practice to key the grid circuits of one of the amplifying stages.

The diode valve (3) and its associated circuits constitute the half-wave rectifier which provides the necessary negative potential required to control the grids of the master valve and the absorber. Condenser (4) is the smoothing condenser for the half-wave rectifier, and resistances (5) and (6)

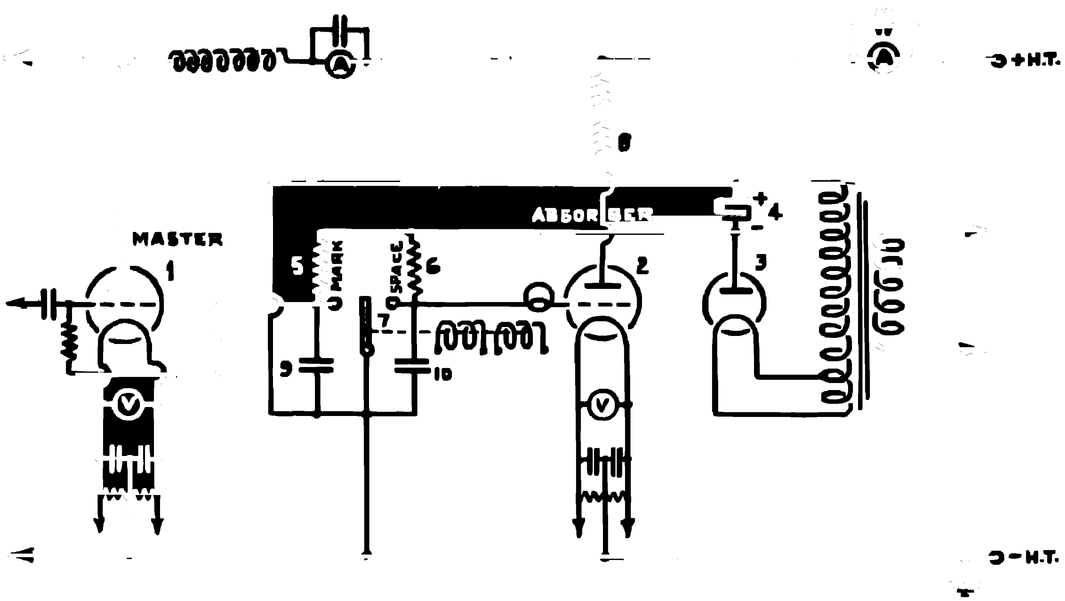


FIG. 57.

are arranged so that the load on the latter remains approximately constant. The operation of the whole arrangement is as follows :—

SPACING POSITION.—In the spacing position, the signalling key (7) is over to the right, and it connects the grid of the absorbing valve direct to its filament. Both electrodes will be at earth potential, current will flow through the valve and the associated resistance (8), and it can be arranged that the power dissipated in this way is equal to that which is used during the mark period. At the same time, the connection between the grid and filament of the master valve is broken, thus leaving it connected to the negative potential supply on the smoothing condenser (4). This negative potential prevents the master valve from oscillating and from taking any anode current. Resistance (6) prevents short circuit of the smoothing condenser (4), and constitutes a load equal to that during the mark period. If a short circuit occurred, there would be no negative potential available for shutting down the master valves.

MARKING POSITION.—In the mark position, the key (7) is over to the left and the opposite action occurs. The grid of the master valve is connected in series with the grid leak to its filament, and the grid of the absorbing valve is isolated from its filament and left connected to the negative supply on the smoothing condenser (4). The master valve oscillates and drives the power amplifying stages which follow it ; during this period, the absorber valve is prevented from taking any anode current. Resistance (5) prevents short circuit of the smoothing condenser. Since its value is the same as that of resistance (6), the small rectifying valve (3) is run under constant load conditions, the two resistances being to this valve what the master and absorber valves are to the main rectifier.

The two condensers (9) and (10), are connected across the key contacts to prevent sparking.

In one Service case, satisfactory signalling was obtained by using a negative potential of the order of 600 volts, and a small key with a break of approximately 1/16th inch. With grid signalling, the speed of operation is chiefly determined by the inertia of the key ; in practice, with keys which can handle bias voltages suitable for shutting down a transmitter employing (say) 10 kW. or more, the operating speed limit is of the order of 150 words per minute.

For very high speeds, up to about 200 words per minute, it is necessary to make smaller both the key and the power which it has to control. This may mean that the bias voltage controlled by

the key may not be sufficient to provide direct control of the grid of the master valve. In that case, some form of valve relay, or D.C. amplifier, must be interposed between the key and the circuits which it has to control. Fig. 58 represents a possible arrangement. The signalling key (7) supplies a small negative bias potential to the grids of valves (1) and (2) during the space and mark periods respectively. Resistances (3) and (4) constitute the anode impedances of these two valves; when there is no negative bias on their grids, current will be passed and a P.D. will be developed across the resistance in the circuit of the valve which is passing current. An amplified negative bias potential is thus available to control the grid of the master valve (5) during the space period, and that of the absorber valve (6) during the mark period. Using a circuit of this kind, the key may be made very small and the gaps between contacts may be reduced almost to the thickness of a sheet of paper.

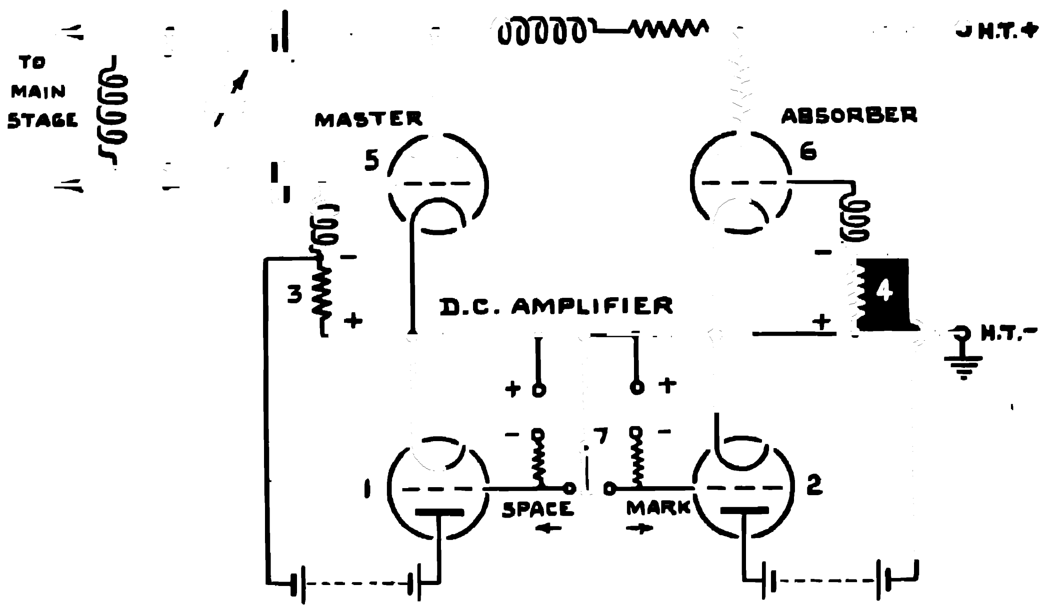


FIG. 58.

60. **V.H/F Oscillations; Applications.**—Frequencies higher than about 25,000 kc/s. are of little use for reliable long distance signalling, since beyond this limit it is unusual for the indirect waves to be sufficiently refracted by the ionosphere to make them return to earth (P.10). Notwithstanding this disability, V.H/F waves are finding an increasing number of applications, perhaps especially due to the new **acorn type** of valve which has made it possible to design receivers capable of working satisfactorily at frequencies above about 60,000 kc/s. Among the applications may be included :—

(*a*) Short range communication, and for all cases where it is desired to restrict a signal, and consequently the interference, to a limited area.
(*b*) Television; the excessive width of the side bands makes it necessary to restrict television to the higher frequencies, in order to avoid the great interference which would result if a carrier wave of lower frequency were used.
(*c*) Radio beacons for the assistance of aircraft, and to facilitate blind landing on aerodromes.
(*d*) Medical applications, including the production of local heating.
(*e*) A number of different laboratory uses.

In the Service, all wave frequencies higher than 30,000 kc/s. receive the title V.H/F. Outside the Service many classifications are found; waves of length between 10 metres and 1 centimetre

(30,000 kc/s. to 30,000 megacycles), are often called "ultra short waves," and waves of length below 1 metre (300 Mc/s.) are usually referred to as "micro waves."

It is proposed to consider briefly the production of V.H/F oscillations.

61. **Production of V.H/F Oscillations by Normal Valve Circuits.**—With suitable care and design, the generation of V.H/F oscillations may be carried out by normal valve circuits up to a frequency limit of about 500,000 kc/s. (60 cms.), the oscillators usually delivering their output directly to the radiating system. At these frequencies, it is difficult to design crystal control generators, and master oscillators with power amplifiers are seldom used above a frequency of about 30,000 kc/s., but, with care, may be used up to 60,000 kc/s.

Fig. 53 represents the circuit details of a transmitter suitable for the production of oscillations up to about 75,000 kc/s.; its basic electrical principles have already been discussed. At these frequencies, the useful range will depend partly on the altitude of the transmitter. Receipt of its transmissions is usually limited to an "optical range" and may vary from (say) 40 miles to several hundred miles.

There are very definite limitations to the highest frequency which can be produced by a normal valve oscillator. These arise from the finite values of the electrode lead inductances and the inter-electrode capacities, and on the finite transit time for the passage of electrons across the valve. At a frequency of about 30 Mc/s. (10 metres), the period of oscillation may be comparable with the transit time of the electrons, a fact which has the effect of upsetting the phase relationships necessary for the maintenance of oscillations.

There are two well-known methods for the production of V.H/F oscillations which do not involve the use of normal valve circuits. These are respectively the Barkhausen-Kurz oscillator, and the Magnetron.

The Barkhausen oscillator is capable of producing low power waves of length down to about 10 centimetres (3,000 Mc/s.).

The Magnetron can produce oscillations covering a wide frequency range, both L/F and H/F, and has been used to produce a V.H/F oscillation, under laboratory conditions, corresponding to a wave as short as 1 centimetre. The Magnetron is the more efficient oscillator, and for frequencies corresponding to waves below 1·0 metres it is preferable to the triode.

62. **The Barkhausen-Kurz Oscillator; Electronic Oscillations.**—Barkhausen and Kurz, two German engineers, employed a triode valve and the circuit shown in Fig. 59. The grid of the triode is kept at a high positive potential, about 150 volts, and the anode is maintained either at zero or a small negative potential with respect to the filament. The external tuned circuit consists of a Lecher wire system connected between grid and anode, the supply voltage being fed into the system at a voltage node through suitable radio-frequency chokes.

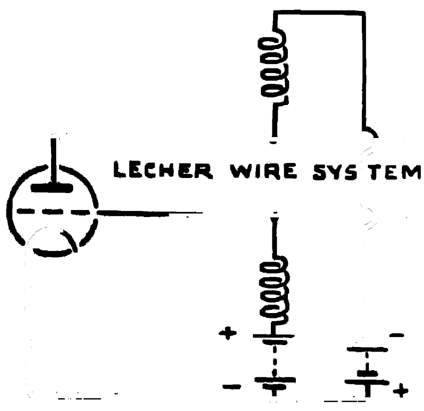


FIG. 59.

Electrons emitted from the filament are accelerated towards the positive grid, through which some of them pass. After passing through the grid, the electrons are subject to a retarding electric field, and, in due course, they are turned back towards the grid. On arrival at the grid they may again pass through it, passing on to the neighbourhood of the filament before the motion is reversed. The electrons may oscillate in this way several times about the grid before being finally drawn to it. Experimentally it was found that these electronic oscillations are capable of maintaining oscillations in a tuned circuit

connected between anode and grid. The frequency generated is therefore very high and depends upon the speed of the electrons, a quantity which is determined primarily by the potentials on the anode and the grid.

Oscillations have been produced in this way for wavelengths ranging from 10 centimetres (3,000 Mc/s.) to a few metres. The worst fault of the B-K oscillator is that the heavy electron bombardment of the grid may sometimes release traces of gas, and make the valve soft. Moreover, the oscillations tend to peak in value at the frequency at which the spiral grid is one full wave in length.

63. **The Magnetron ; Early Form and Basic Theory.**—The first mention of the magnetron was made by A. W. Hull in 1921, when an account was published of a cylindrical diode with a uniform magnetic field in the direction of the electrode axis, *i.e.*, parallel to the axis of the filament. In its first form, the magnetic field was obtained by surrounding the diode with a solenoid, as shown diagrammatically in Fig. 60 (*a*).

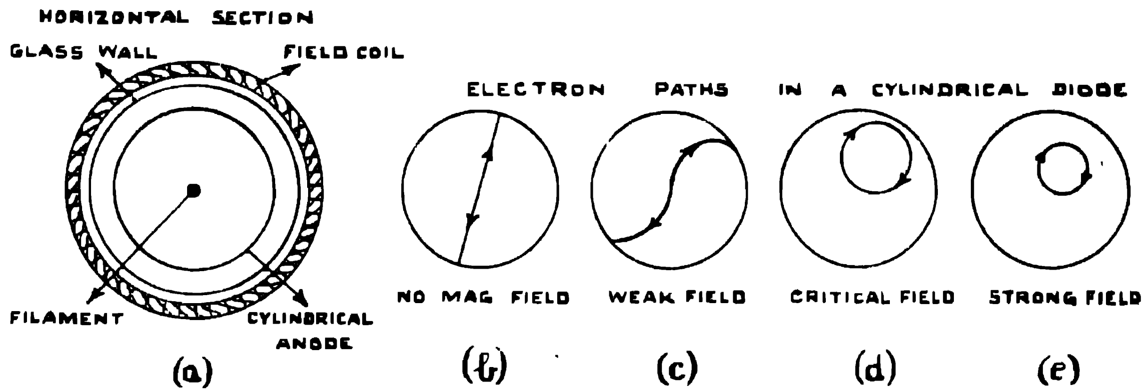


FIG. 60.

The behaviour of the electrons within that valve may be studied with the help of Fig. 60 (*b*), (*c*), (*d*) and (*e*). The anode is maintained at a positive potential with reference to the filament.

With no magnetic field, electrons will travel radially under the influence of the electric field. With a weak magnetic field, the electrons will be subject to a deflecting force acting at right angles both to the magnetic field and to the direction of motion of the electrons at any instant. It will be recalled that an electron in motion has all the properties of a current bearing conductor, and the direction of movement (or deflection) of a conductor when placed in a magnetic field is given, simply, by applying Fleming's left-hand rule. A deflecting force which is at right angles to the direction of movement of an electron must produce circular motion, the electron is like a stone at the end of a sling. For this reason, electrons traverse circular arcs between the filament and the anode. If the field strength (expressed in Gauss) is increased further, the radii of the circular electron paths will decrease, until, with a certain critical value of the field, the electrons fail to reach the anode at all and follow circular paths making grazing incidence with the anode. For still stronger fields, the radii of the electron paths become still smaller, as shown in Fig. 60 (*e*).

Fig. 61 represents two curves of the family of static characteristics of such a valve. As would be expected from the above theory, for a given anode voltage, the anode current remains steady over a range of values of the magnetic field, but experiences a sharp "cut-off" when the critical value of the field is reached. Theoretically, the anode current should fall instantly to zero ; in practice this is not the case, and curves of the form of those in Fig. 61 are obtained. There are various possible ways of accounting for this slight abnormality. It is not, for example, strictly true to say that all electrons leave the filament with the same velocity, or attain the

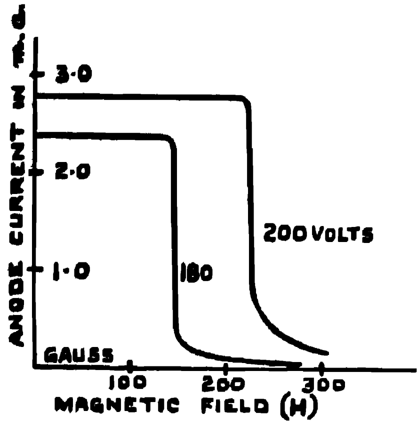


FIG. 61.

same velocity after travelling to the anode. Moreover, the magnetic field is not uniform throughout the length of the anode, and electrons escape at the ends.

The form of the characteristics shown in Fig. 61 suggests that **the magnetic field could be used to control the anode current in the same way that the grid voltage does in the case of the triode oscillator.** This principle was actually used in the first magnetron oscillator; in that form it can, however, only be used for the production of low frequency oscillations. In the modern form of the magnetron, a constant magnetic field is used to control the *direction* rather than the *magnitude* of the anode current.

The frequency limitations and relative inefficiency of the first magnetron oscillator made it a poor competitor to the triode oscillator, until it was found possible in 1924 to use it in the production of H/F oscillations. In 1929, the magnetron began to take its present form in which the anode is divided into two or more equal segments separated by narrow air gaps; first mention of this was made by Okabe in Japan. The most common form has two anode segments, and is usually called the "**split anode magnetron.**" It is this splitting of the anode which has increased the efficiency of the magnetron to such an extent that it is now preferable to the triode for wavelengths below 1·0 metre.

★64. MATHEMATICAL NOTE ON CIRCULAR ELECTRON ORBITS.—Ampère's law (Vol. I) state that the force on a current bearing conductor placed in a magnetic field of flux density B is given by

$$F = BI \,.\, \delta l \,.\, \sin\theta.$$

An electron in motion constitutes a current, and, in this case, the force on one electron having a charge e, may be expressed by

$$F = Bev \quad \ldots\ldots \quad \text{where } v = \text{velocity of the electron.}$$

Hence $\quad F = Hev \quad \ldots\ldots\ldots \quad (1)$

since $\quad B = H$ for air (where H is the field strength).

By Fleming's rule, this force acts at right angles both to the direction of electron movement and to the magnetic field; circular motion therefore results, as shown in Fig. 59.

The electron velocity remains unaltered by the magnetic field. Its magnitude is determined by the electric field, since the KINETIC ENERGY GAINED by an electron of mass m in moving from the filament to a point P where the potential is V_0 is EQUAL TO THE WORK DONE BY THE ELECTRIC FIELD.

$$\therefore \quad \tfrac{1}{2}mv^2 = eV_0 \quad \ldots\ldots\ldots \quad (2)$$

The electric field between a thin filament and a concentric anode is very strong in the neighbourhood of the filament, and very weak elsewhere. As a first approximation, we may assume that all of the voltage drop V_0 occurs near the filament, and that, in the region from there to the anode, the electrons move with constant velocity v given by (2).

Under these conditions the magnetic field makes the electrons move in circular orbits of radius "r" given by

$$\frac{mv^2}{r} = Hev$$

$$\text{or} \quad r = \frac{mv}{He} = \frac{1}{H}\sqrt{\frac{2mV_0}{e}} \quad \ldots\ldots\ldots \quad \text{from (2)}$$

$$\text{or} \quad r \propto \frac{\sqrt{V_0}}{H} \quad \ldots\ldots\ldots \quad (3)$$

Also, if the period for one circle is T,

$$2\pi r = v1. \quad \therefore \quad T = \frac{2\pi r}{v} = \frac{2\pi m.}{He} \quad \ldots\ldots\ldots \quad (4)$$

Also, from (3), $H = \frac{1}{r}\sqrt{\frac{2mV_0}{e}}$ and the " critical value " of H can thus be shown to be approximately given by

$$H = \frac{\sqrt{180V_0}}{d} \quad \ldots\ldots\ldots \text{ in Gauss, where } V_0 = \text{anode volts, where } d = \text{anode diameter (cms.).}$$

65. **The Modern Split Anode Magnetron.**—Fig. 62 (*a*) represents the valve in its simplest form. Instead of providing the magnetic field by means of a solenoid, it is now usual to dispose the electrodes so that the field may conveniently be provided by means of an external electromagnet, as shown in Fig. 64; in certain cases, the field may be provided by permanent magnets.

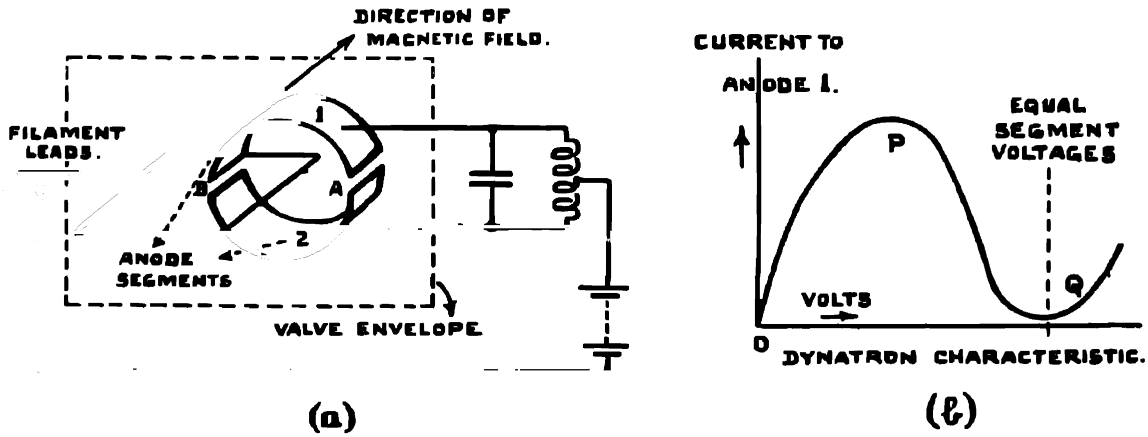


FIG. 62.

Under suitable conditions, the *basic* magnetron *circuit* of Fig. 62 (*a*) is capable of providing oscillations covering an unbroken frequency band extending from low frequencies of the order of 100 kc/s. (3,000 metres) to 30,000 Mc/s. (1 centimetre); naturally, it would need several magnetrons to cover this range.

It is, however, customary to distinguish an essential difference between the mechanism of oscillations in the frequency band below 30,000 kc/s., and that involved in the production of oscillations in the band above 500 Mc/s. (60 centimetres). The former are usually called "**dynatron**" **oscillations**, and the latter "**electronic**" or "**resonance**" **oscillations**; the word dynatron is used since it is now well known to refer to a valve—such as a screened grid valve—having a static characteristic over a part of which the slope is negative, corresponding to negative resistance. Between these bands there is a range of intermediate frequencies of great practical importance. Although this distinction is made between the two types of oscillation, it is, however, probable that all oscillations may be attributed to the same cause, namely, the PRODUCTION OF NEGATIVE RESISTANCE.

In this instance the two extreme types of oscillation will be briefly described, although the mos important frequencies at present fall within the intermediate band.

NEGATIVE RESISTANCE—DYNATRON OSCILLATIONS.—Fig. 62 (*b*) represents the static characteristic of a split anode type magnetron valve, in which one anode segment (say) No. 2 is held at a constant potential with reference to the filament, the magnetic field strength being above the critical value (Fig. 60 (*d*)). If the potential of the other segment is varied and the anode current to it measured, a characteristic will be obtained having the form indicated.

When the voltage on segment No. 1 is very small, it receives very few electrons but a large current flows to the other segment, as shown in Fig. 63 (*a*). As the voltage increases, electrons are drawn in increasing numbers (point P), but when the voltages on the two segments approach equality,

the action of the critical magnetic field becomes fully operative, and the electrons follow circular paths missing both anode segments, Fig. 63 (*c*). The latter condition corresponds to the point Q, and it is clear that the portion of the curve PQ represents a negative resistance, since the current decreases as the voltage increases.

By connecting a symmetrical push-pull circuit between the two segments, as shown in Fig. 62 (*a*), it is possible to make use of this negative resistance and supply energy to the circuit.

In order to see how an oscillation can be maintained, once it is started, it is necessary to have some theory giving an explanation of how the segment at a lower potential $V - \delta V$ can attract a greater number of electrons than the segment at a higher potential $V + \delta V$. A simple qualitative explanation has been produced by Gill and others, and is represented diagrammatically in Fig. 63 (*d*). The electric field is radial and symmetrical when the two segments are at the same potential V. If an oscillation commences, the segments take up instantaneous potential $V + \delta V$ and $V - \delta V$; since the two segments are then at different potentials, there will be an electric field between them, which superimposes itself on the original radial field. This produces a distortion of the electric field which is held to be most serious in the neighbourhood of the gaps A and B. The field near A, which was originally radial, is turned through a small angle and it can be shown that an electron which would normally have followed a circular path and missed both anodes, receives a deflection

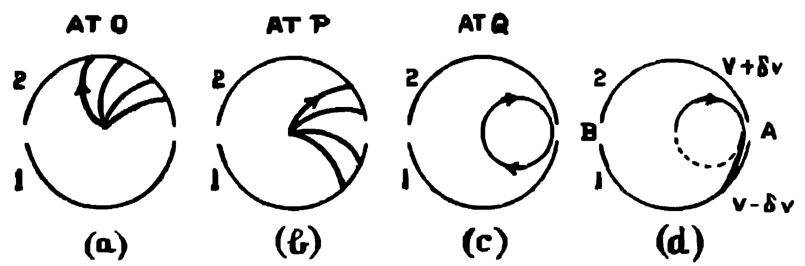


FIG. 63.

near A which sends it to the segment at the lower potential. By similar reasoning, depending upon the distortion of the field, it can be shown that an electron which was near the top of its path and close to the segment at the higher potential, receives a deflection which takes it further away.

The essential characteristic of these simple dynatronic oscillations is that **the frequency generated is substantially dependent upon the natural frequency of the symmetrical tuned circuit attached to the valve.** For a given voltage the best working value of magnetic field is the same for all wave frequencies.

RESONANCE OR V.H/F ELECTRONIC OSCILLATIONS.—Although a magnetron is capable of giving oscillations covering an unbroken frequency band, it is found that at frequencies corresponding to waves of length about 75 centimetres (and shorter), for a fixed anode voltage, the system is no longer independent of the value of the magnetic field H. For a fixed value of anode voltage, as the magnetic field is gradually increased it is found that the amplitude of oscillations depends on H, the system oscillating strongly at certain frequencies corresponding to particular values of H. The essential nature of these oscillations suggests a process something akin to resonance, and the term "**resonance oscillation**" is applied by Gill, or "**selective negative resistance**" by Megaw.

Experimentally it is found that for a given ratio of V and H, there is some frequency at which the system oscillates best. This implies that the frequency is proportional to the ratio $\frac{V}{H}$, and expressed by the relation $f = K\frac{V}{H}$, where K is a constant.

The above relation has not yet been satisfactorily deduced from theoretical considerations; for a given frequency it gives a straight line relation between V and H.

The theory of the mechanism of these resonant oscillations is at present in an incomplete state. It is, however, thought to depend upon the production of a form of negative resistance, in which the electrons may have made *several spiral orbits* within the valve, being deflected in the same direction by successive anode segments, before eventually landing on an anode segment at a lower instantaneous potential. If any such process is occurring, it can easily be understood that the frequency of a strongly sustained oscillation is intimately connected with the time of oscillation of a single electron, and hence with V and H.

The requisite condition would seem to be that an electron should complete a whole number of orbits in the time interval between successive gaps A and B. That time interval should then correspond to half the periodic time of the generated oscillation, since an electron which does not "land" on the lower segment at A *must not* land on the upper segment at B, if the necessary conditions of negative resistance are to be maintained.

This theory is referred to by Gill as the "**precession of the electrons**" within the valve. Fairly complete confirmation of it has been obtained by the use of a special magnetron having four anode segments, the alternate segments being electrically connected; the special valve may be called the **2-section split anode magnetron.** Working both type of magnetron under the same conditions, Gill found that the frequency generated by the 4-segment type was approximately twice that produced by the 2-segment type.

Further very elegant experimental evidence has been produced by Kilgore demonstrating the existence of spiral electron orbits in a magnetron. By introducing a little argon into the special magnetron operated under static conditions, the electron orbits were made *visible* and actually photographed.

A simple extension of this theory shows that if the magnetic field is inclined to the axis of the electrodes, the resultant motion of the electrons becomes helical instead of spiral. In practice, this may be used as one of the variable factors controlling the operation of the valve.

More recently, Megaw and others have restricted the use of the term electronic oscillations to oscillations of the highest frequency only. They are the ones which are obtained by means of

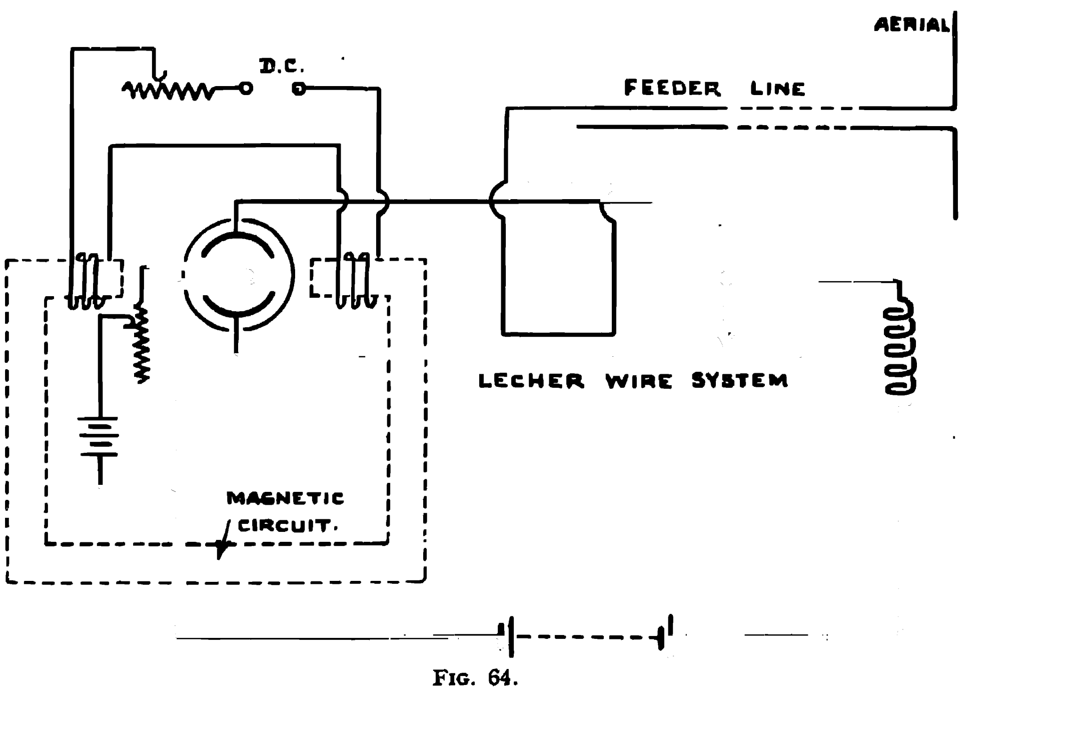


FIG. 64.

resonance with the frequency at which the electrons spiral around the lines of magnetic force. Megaw has shown that these electronic oscillations are only possible when the emission from the filament is small, implying a negligible space charge; he also conludes that the maintenance of electronic oscillations in a magnetron with the magnetic field exactly parallel to the filament (and perfectly symmetrical electrodes) is impossible. In that case, it appears that oscillations are contingent upon a slight inclination of the magnetic field to the axis of the electrodes.

INTERMEDIATE RANGE OSCILLATIONS.—According to Megaw, oscillations covering the band extending from about 30,000 kc/s. to 600 Mc/s. are intermediate in nature between dynatron oscillations and resonance oscillations. It may be considered that these oscillations are essentially dynatron oscillations in which the transit time of the electrons exercises an increasing influence of the frequency generated as the frequency is raised. They are the frequencies for which it is not true to say that the generated frequency is mainly determined by the constants of the tuned circuit attached to the anode segments; the frequency limits are not fixed but depend on the valve dimensions.

Fig. 64 represents a complete magnetron circuit, essentially suitable for the production of frequencies with any range. The anode segments are connected to a lecher wire system which is fed through a suitable H/F choke attached to a moving slider. A dipole aerial is shown coupled to the lecher wires. For lower frequencies it would be necessary to replace the lecher wire system by a tuned circuit, as in Fig. 62.

One of the most serious difficulties attached to the use of a magnetron is that of modulating the oscillation. Various methods have been tried including a variation of the magnetic field, and the use of a grid. Considerable success has attended the use of grid wires, and these have been found to have the additional virtue of protecting the filament, in some measure, from bombardment by returning electrons. The problem cannot yet be considered to have been satisfactorily solved.

SECTION " K."

VALVE TRANSMITTERS.
EXAMINATION QUESTIONS.

1. In its simplest form, sketch a self-oscillatory valve oscillator employing a Hartley circuit and parallel feed. By means of curves, indicate the essential phase relationships existing between the anode oscillatory potential, the grid oscillatory potential, and the oscillatory anode current. In this way, explain why the circuit is self-oscillatory.

(Qual., Lieutenant (S.), 1935.)

2. (*a*) Outline the evolution of the service valve transmitter circuit known as the " divided circuit." Show, with the aid of diagrams, how the correct phase relationships are obtained for the maintenance of oscillations.

(*b*) What is the function of the anode tapping point and how may the correct setting be derived from theoretical considerations?

(Qual., Lieutenant (S.), 1936.)

3. (*a*) What are the main causes of frequency variation in self-oscillatory transmitters?

(*b*) How may the piezo-electric effect of quartz be utilised to stabilise the frequency of a transmitter?

(Qual., Lieutenant (S.), 1936.)

4. " The main stage of a master controlled transmitter is similar in principle to the output stage of a receiver." Justify this statement by a description of the action of any such transmitter you have met, giving the circuit diagram.

(Qual., Lieutenant (S.), 1933.)

5. What are the objects to be attained when neutralising a master controlled transmitter, and what are the possible effects of bad neutralisation? Outline the practical procedure to be adopted when neutralising, and mention the difficulties that arise in neutralising an H/F circuit.

(Qual., Lieutenant (S.), 1933.)

6 Sketch a valve transmitting circuit which has parallel feed, tuned circuit between anode and filament, direct inductive grid excitation, and anode at high oscillatory potential. Describe its action from the point of view of maintenance of oscillations, and comment briefly on the arrangements for securing high efficiency.

(Qual., Lieutenant (S.), 1932.)

7. Discuss the advantages and disadvantages of mutual aerial excitation, and explain the phenomenon of frequency jump.

(Qual., Lieutenant (S.), 1932.)

8. What are the reasons for the use of piezo-electric crystals for frequency control? What precautions are necessary if the best results are to be obtained? What approximately are the highest frequencies for which quartz crystals can be manufactured and used? How can frequencies above this limit be controlled?

(C. & G., Final, 1934.)

9. Describe how, by employing a valve with reaction, the high frequency resistance of an oscillatory circuit may be reduced. Find an expression connecting the oscillatory current with the applied potential difference in terms of the mutual inductance of the reaction coil, and the circuit and valve constants, and hence find the mutual inductance necessary for self-oscillation to commence.

(C. & G., Final, 1926.)

Explain how a three-electrode valve may be used to produce high frequency alternating currents. Discuss the design of the circuit arrangements.

(I.E.E., October, 1927.)

Section "K."

11. Show that, if an oscillating valve circuit is arranged so that the maximum power output is obtained, the efficiency (neglecting the power used in heating the filament) is only 50 per cent. Explain how the efficiency may be increased, and state the attendant disadvantages.

(L.U., 1930.)

12. Describe, with schematic diagrams, three different circuits with which a three-electrode valve can be used to generate oscillations.

(C. & G., 1., 1930.)

13. Find expressions for the condition for maintenance, and for the frequency, in a valve oscillator with the tuned circuit connected to the anode. Show all the voltages and currents involved in a vector diagram, and indicate how the frequency of the oscillation departs from the resonance frequency of the oscillatory circuit. Suggest a method of eliminating this departure.

(L.U., 1932.)

14. A tuned circuit of $L = 500\mu H.$, $R = 20$ ohms, and $C = 300\mu\mu F.$, is connected across the grid and filament of a triode of $\mu = 10$ and $r_a = 50{,}000$. In the anode circuit is a coil $L_a = 100\mu H.$, coupled with the inductance L by a mutual inductance of $20\mu H.$ Estimate the alteration the presence of the valve makes in the effective resistance of the tuned circuit. (Result : $13\frac{1}{3}$ ohms.)

(I.E.E., May, 1934.)

RADIO-TELEPHONY. SOUND REPRODUCTION.

1. **Essential Problems and Features of R/T.**—Radio-telephony is the name given to the transmission by electromagnetic waves of the various audio-frequency sounds which compose speech, music and the noises of everyday life. It bears the same relation to W/T as line telephony does to line telegraphy, and possesses some of the same relative advantages and disadvantages.

From the consideration which has already been given to the production of modulated wave forms (K.52), it will be appreciated that the transmission of speech and music involves no essentially new problem. The production of I.C.W. wave forms having an audio-frequency modulation, and in some cases a supersonic modulation, has been seen to be a comparatively simple matter. The ordinary I.C.W. transmitter, in effect, arranges to radiate a constant note, say a 1,000-cycle note, and the sound of the latter is broken up into the shorts and longs of the morse code by means of a key. In radio-telephony, the modulation impressed on the C W oscillator at the transmitter must not correspond to a simple 1,000-cycle note, but to the sum total of all of the component frequencies in the sound which it is desired to transmit. Essentially, the same methods of modulating, which have already been discussed in the case of I.C.W., apply equally to R/T.

From this it should be clear that R/T became possible when the C.W. oscillator was first produced; the development of R/T has, therefore, followed very closely the course of development of the valve transmitter itself. In transmitting R/T the unmodulated wave—or CARRIER WAVE—is often radiated continuously, being intermittently modulated by the various A/F sounds. A microphone and microphone amplifier will virtually replace the morse key employed in ordinary Wireless telegraphy.

In R/T reception certain features characterise the receiving apparatus, which, in the main, is the same as that employed in W/T reception.

The receiving apparatus must use telephones or a loudspeaker as the final indicator of the signal, since no automatic recording device giving a result which can be interpreted by the eye is yet possible; automatic recording for aural reproduction is widely used.

Heterodyne reception obviously need not be used, but if it is employed, the heterodyne frequency must be adjusted to the centre of the "dead space," or an undesired beat note will be heard. The use of heterodyne reception will always introduce some distortion; it is seldom used, but is sometimes of value in increasing the sensitivity of the detector when very weak signals are being received. The above remarks should not be confused with "superheterodyne reception" which is quite satisfactory for R/T, and is the principle of operation of the majority of R/T sets employed for the reception of broadcast programmes.

The modern broadcast receiver has many complicating refinements involving auxiliary controls which may not, in general, be found in a receiver designed especially for telegraphy. These refinements include automatic gain control (A.G.C.), automatic tuning control (A.T.C.), tone control, variable selectivity, volume expansion circuits, tuning indicators, second channel interference suppressors, etc.

Many precautions must be taken to prevent distortion of the audio-frequency wave forms if it is hoped to produce a result which is a faithful reproduction of the original sound. Sources of distortion exist both at the transmitter and at the receiver. At the transmitter it may arise due to the use of inferior microphones or incorrect control of the modulation. At the receiver it may be due to the use of circuits which are too sharply tuned, possibly the result of too much "reaction." The effect of the latter is to attenuate some of the higher modulating frequencies, a process which is known as "side band cutting"; it is overcome in practice by the use of "band-pass" circuits. Other causes of distortion at the receiver may include faulty detector adjustments, over-loaded valves in the A/F stages, inferior iron-cored chokes and transformers, inferior reproducing devices, loud speakers, etc.

It should be noted that side band cutting may also originate at the transmitter; amplifying stages following the modulator should possess band-pass characteristics.

2. **Advantages and disadvantages of R/T.**

ADVANTAGES.

(*a*) All the normal advantages of oral over written communication, *e.g.*, actual saving in time and the reduction of traffic that arises by the adjustment of minor points without the necessity of voluminous correspondence.

(*b*) The ability to make and read morse is not required.

DISADVANTAGES.

(*a*) Where a message has to be written down, R/T is slower than W/T; it compares unfavourably with high speed morse which gives a typed record.

(*b*) Coded messages are difficult to pass, owing to the risk of phonetic errors, and plain language involves the risk of interception unless some "privacy method" of communication is employed; variety of "accent" adds to the difficulty.

(*c*) The interference caused by R/T is worse than that caused by W/T. In the case of the B.B.C. transmitters, it will be shown that the interference covers a band approximately 16 kc/s. in width.

(*d*) "Listening-through" is not possible with R/T unless the "single side band" system is employed.

3. **Speech and Music.**—Speech and music consist of highly complex sound waves, *i.e.*, waves containing many different frequencies and rapidly-varying amplitudes. Examples of such wave forms are shown in Figs. 1 and 2. Before proceeding to an analysis of the composition of such waves and their effect on the human ear, so that the points involved in their transmission by R/T without serious distortion may be clearly grasped, it is convenient to define the following terms associated with speech and music.

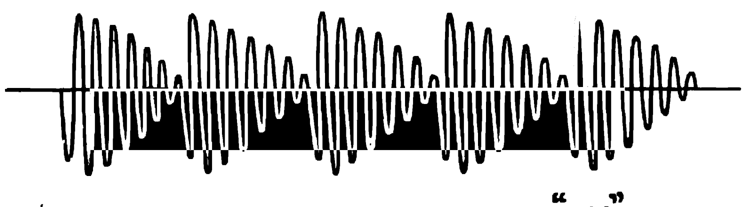
SOUND INTENSITY is a measure of loudness and depends on the amplitude of the sound waves.

Pitch may best be described by comparing it with colour in vision. It is a physiological effect which depends on the frequencies in the speech or music. In the same way as the colours red and blue describe light in which low and high visual frequencies respectively predominate, so the terms "low pitch" and "high pitch" describe speech or music in which low and high audio-frequencies respectively predominate.

The term "pitch" is also used in another sense to denote the standard frequencies used in music. For example, in "concert pitch" the note "middle C" has a frequency of 273 cycles per second; in French standard pitch the note "middle C" has a frequency of 261 cycles per second, while for scientific purposes this note is taken to have a frequency of 256 cycles per second.

A pure tone is a pure sinusoidal wave of constant frequency and of sensibly constant amplitude. This rarely, if ever, occurs in practice, one of the nearest approaches to it being the note produced by a tuning fork.

A complex tone is a combination of waves of several frequencies which may be of constant or varying amplitude. A wave-form at a single frequency which varies rapidly in amplitude is also equivalent to a complex tone. Any complex tone can be analysed into component "pure tones"



WAVE FORM OF VOWEL SOUND "AH!"

Fig. 1

over a short period. The lowest component pure tone is known as the "fundamental," and the remaining tones as "overtones." The fundamental tone is usually, but not always, the strongest, and therefore predominates. The fundamental tone usually gives a complex tone its characteristic pitch (determines the note), while the overtones are responsible for the characteristic "timbre" or "tone quality" (*see* below). The component pure tones which go to form a complex tone are often called "partials."

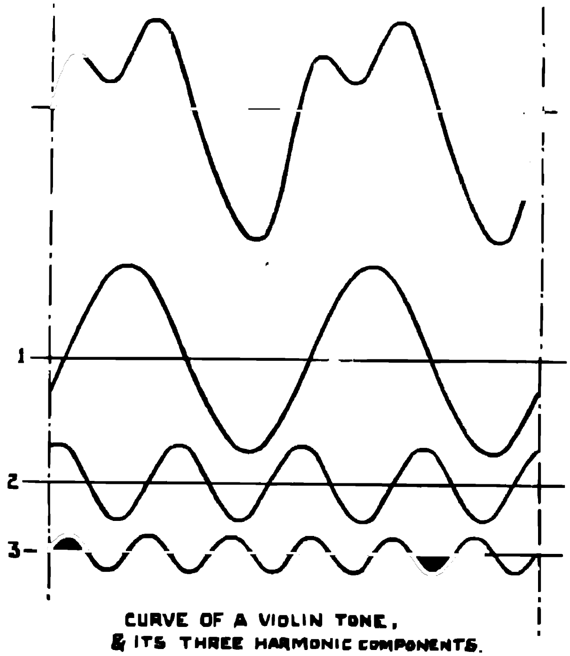


FIG. 2.

Harmonics.—Overtones whose frequencies are exact multiples of that of the fundamental tone are called "harmonics." The fundamental is also called the first harmonic. The wave whose frequency is twice that of the fundamental is known as the second harmonic. The third harmonic has a frequency three times that of the fundamental, and so on. Fig. 3 shows a stretched string vibrating freely at its fundamental frequency and also at its second, third and fifth harmonics.

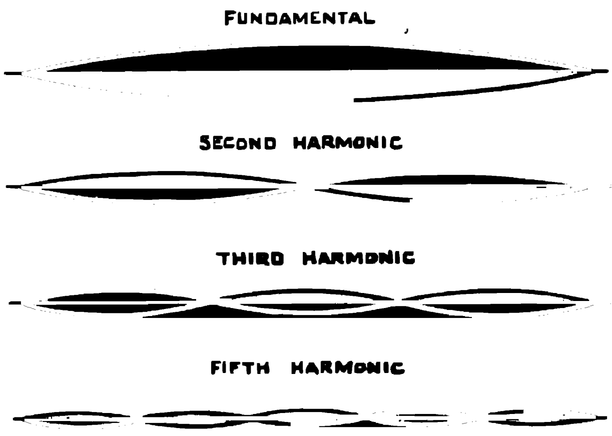


FIG. 3.

The second harmonic of any fundamental frequency is an "octave" above the fundamental; the third harmonic is a "fifth" above the second, and so on.

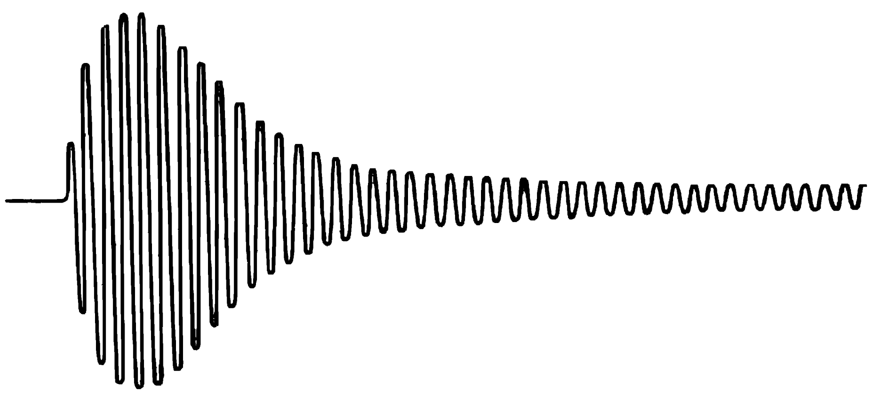
Timbre or Tone Quality.—Notes of the same fundamental frequency produced on different instruments give rise to very different sound sensations when they impinge on the human ear. For example, if "middle C" (256 cycles per second) is played on the flute and the same note is played on a violin, although the note will appear the same, the tone quality is entirely different, and it is possible to tell which instrument has produced it. This is due to the fact that the overtones accompanying the fundamental frequency differ according to the instrument. The note of a tuning fork is almost entirely free from overtones, and the note sounds "pure." A flute produces a note which contains a few weak overtones. A violin produces much stronger overtones, and the note is "harder" than that of a flute. Reed instruments produce very strong overtones, so much so that in some cases it is difficult to distinguish the fundamental note.

In many cases of complex sound waves, it is possible to suppress the fundamental frequency entirely and yet to hear it. This is due to a rectifying action in the ear which produces a low frequency from the combination of two overtones of much higher frequency. This effect can be observed in portable gramophones and small horn-type loud speakers, where the lowest note of the piano (of fundamental frequency about 50 cycles per second) can be heard, although the instrument is incapable of reproducing to any extent frequencies below about 200 cycles per second.

Hence it will be seen that "tone quality" lies almost entirely in the overtones, and any reproducing system must reproduce these as faithfully as the fundamental if the tone quality is to be distinguishable. Normally, for good (but not perfect) reproduction, the general tone quality of any note will be distinguishable if only the overtones up to twice the frequency of the fundamental (*i.e.*, the second harmonic) are uniformly reproduced.

4. **Transients** (known in musical parlance as "attack") are produced by rapid changes in sound intensity or sudden impulses of a non-periodic nature. Violent transients appear to the human ear as **noise** without any definite note (*e.g.*, pistol shots, slamming doors, etc.). This effect can be compared to that of atmospherics, which, being heavily damped, exert a shock effect on aerials tuned to almost any frequency, and consequently appear to possess no definite periodic frequency of their own.

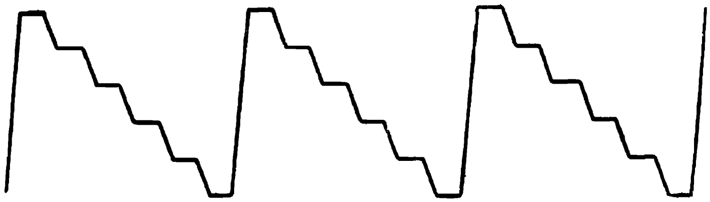
Transient conditions occur when enunciating words with explosive initial consonants, such as *C*at, *B*oy, *T*own, and in the sounds produced by percussion instruments such as cymbals, bells, drums, and, to a certain extent, the piano. Fig. 4 shows the transients produced at the commencement of a note played on a piano.



VIBRATION OF PIANO STRING AT COMMENCEMENT OF NOTE SHEWING TRANSIENT CONDITION AND LARGE INITIAL AMPLITUDE.

FIG. 4.

Other examples are the heavily-damped sound waves caused by hand-clapping and pistol shots. Angular or rectangular wave forms also give rise to them. The wave-form of any stringed instrument played with a bow is usually of an angular nature, as shown in Fig. 5. Transients are among the factors which determine Tone Quality.



HELMHOLTZ'S DIAGRAM OF THE VIBRATIONS OF A VIOLIN STRING UNDER THE ACTION OF A BOW, SHEWING THE ANGULAR WAVE-FORM

FIG. 5.

Transient wave-forms can be analysed (by Fourier's Theorem) into a large or infinite number of periodic sound waves, the frequencies of which vary from zero to infinity. From this it will be seen that any apparatus capable of reproducing perfectly all transient wave-forms must be capable of reproducing uniformly all frequencies from zero to infinity. The converse is also true, namely that any apparatus which will reproduce **uniformly** all frequencies from zero to infinity, will also reproduce all transient wave-forms perfectly.

The human ear, however, can only detect frequencies lying between 20 and 20,000 cycles per second. Consequently, any apparatus which will reproduce uniformly all frequences within this range will also reproduce all transients as perfectly as can be detected by the human ear.

Any resonances in a reproducing system will cause imperfect reproduction of transients. The shock effect of the latter will give rise to disproportionately loud reproduction at any resonant frequency in the audible range, and the transient wave-form will therefore be " coloured " by that frequency.

5. **The Human Ear.**—A simple explanation of the action of the human ear is to consider it as containing a very large number of tightly-stretched "strings," each of which is tuned to a certain frequency. The resonant frequencies of these " strings " are separated by very narrow intervals, and are spread over a frequency range of about 20 cycles per second to 20,000 cycles per second. Each " string " is connected by a nerve to the brain. When any particular " string," *e.g.*, that tuned to 1,000 cycles per second, is set into vibration, the consequent nervous impulse reaching the brain causes a sensation which the brain has learnt to associate with a 1,000-cycle note.

If a pure tone impinges on the human ear, only one " string " (theoretically) is set into vibration. If, however, a complex tone reaches the ear, the complex tone is automatically analysed into its component pure tones (viz., the fundamental tone and the overtones). Each of these component pure tones sets in vibration the corresponding " string " in the ear, and the overall nervous impulse reaching the brain gives rise to the sensation which the brain has learnt to associate with that particular complex tone.

The human ear is not affected by the relative phases of the component frequencies of a complex tone, as would be expected from the above theory. Therefore any reproducing apparatus which does not reproduce all audio frequencies in their correct relative phases may still give distortionless reproduction.

The effect of transients on the ear is to shock some or all of the " strings " in the ear into vibration for a very small space of time. This effect is analogous to that produced by slamming the lid of a piano when the right pedal is pressed, which causes all the piano strings to vibrate. The effect on the brain of the resultant jumble of nervous impulses, is that which it has learnt to associate with a " noise " which has no definite tone or frequency.

6. Requisite Modulating Frequency Limits.—For **perfect** reproduction, all the frequencies which are audible to the human ear should be uniformly reproduced. In practice, however, it will usually be satisfactory to work with a considerably smaller band of frequencies.

If all frequencies below 40 cycles per second are removed, only the very deepest pedal notes of an organ and the fundamental tones of the double bass, bass drum, etc., will be affected. At the present time it is not generally practicable to reproduce frequencies lower than about 40 cycles per second at anything approaching their true relative intensity.

If all frequencies below 300 cycles per second are removed, the general **clarity** of speech or music will not be seriously impaired, but the reproduction will be thin and lacking in body.

If all frequencies above 10,000 cycles per second are removed, only the very finest shades of tone will be affected, and this defect in reproduction will normally be detectable only by those with musically-trained ears. The effect on speech will not be noticeable.

If all frequencies above 5,000 cycles per second are removed, the natural quality of speech and music will be slightly impaired. With a large orchestra, it will not be so easy to pick out the individual instruments, and the finer differences between similar types of instrument will be lost. In the case of speech it will not be so easy to recognise the voice of the operator.

If all frequencies above 3,000 cycles per second are removed, speech will still be intelligible, but will have lost its natural quality. The consonants will be weak, especially the sibilants. Music will be "drummy," and will have lost most of its natural quality.

Thus the frequency bands which should be retained in practice for radio telephony are roughly as follows :—

For intelligible speech	300 to 3,000	cycles/sec.
For speech where special clarity is desired (*e.g.*, for important orders delivered by R/T)	200 to 5,000	,,
For music where **realistic** quality is not essential	150 to 5,000	,,
For music where **realistic** quality is required	50 to 8,000	,,

The latter frequency band (50—8,000) is that covered approximately by the British Broadcasting Corporation's transmitters.

In the case of one of the Service Warning Telephone (Wa/T) outfits—a type of public address equipment—special arrangements are made to introduce attenuation of the low notes in normal speech, relative to the middle and upper frequencies, in order to reduce "boom" and "echo" which are so pronounced in a ship as to mask the intelligibility of speech. These arrangements include a fixed amount of attenuation of speech frequencies below 400 cycles per second, and a variable amount of attenuation of speech frequencies below 800 cycles per second. At the same time, accentuation of the higher frequencies of speech in the band from 2,000 to 6,000 cycles per second is provided to emphasise the consonants and sibilants, and so improve the intelligibility of speech. The latter relatively weak sounds are rapidly lost in the poor acoustic conditions which exist in a ship. (Paragraph 41.)

7. Requirements for R/T and Sound Reproducing Apparatus.—In order that speech or music may be satisfactorily transmitted or received by any apparatus, the latter must be designed to fulfil the following requirements :—

(*a*) All frequencies within the limits specified in the above paragraph must be uniformly reproduced. This implies that the voltage amplification factor (V.A.F.) of the A/F stages must be the same for each of the frequencies in the sound which is being amplified. If this is not the case **frequency distortion** will be the result.

In general, this is avoided by designing an output impedance which is independent of the frequency, thereby avoiding all resonances.

The relative intensity of the reproduction and the original must be constant for all sound intensities. This implies that the V.A.F. of the A/F stages must be

independent of the amplitude of the input signal. If this is not the case **amplitude distortion** is the result (*cf.* Appendix A.9).

It is possible to deduce all forms of distortion to one or the other or a combination of the above types.

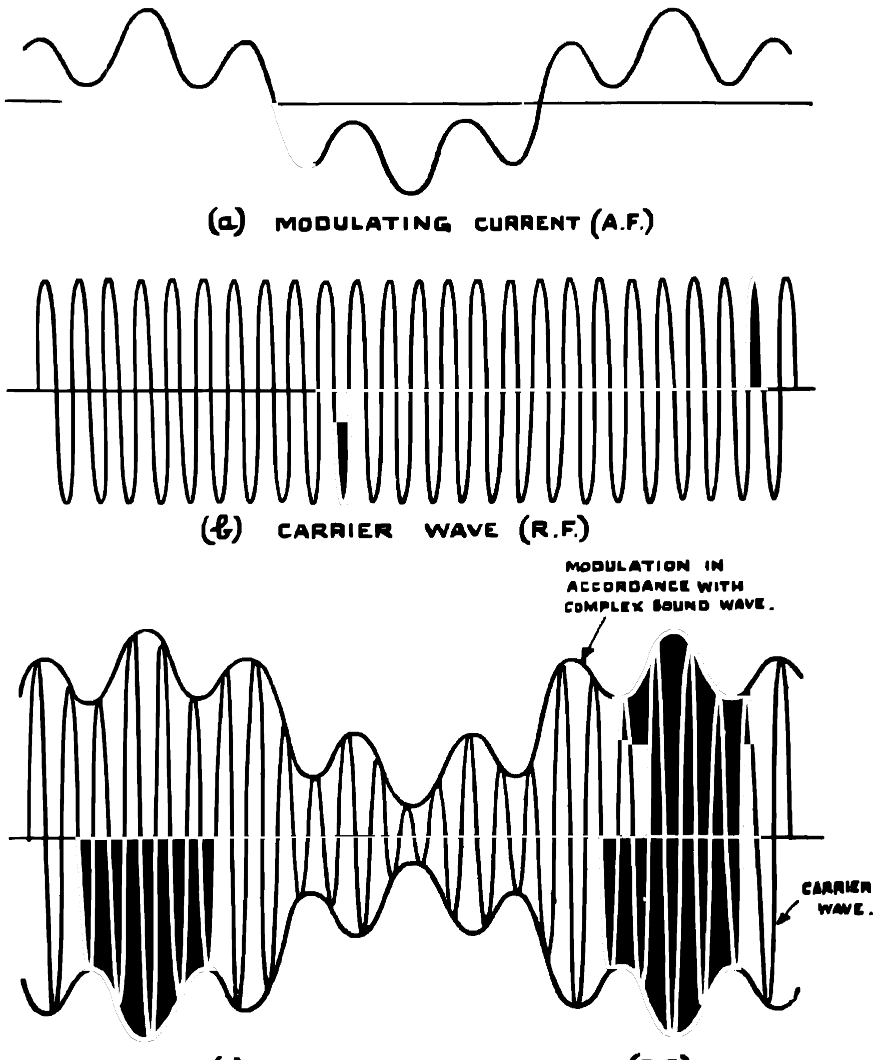
8. **Essential Processes in Radio Telephony.**—There are four processes which must always be employed in transmission and reception of speech or music. These are :—

(*a*) Conversion of sound vibrations in the air into corresponding audio-frequency oscillatory currents and voltages. This is effected by a microphone.

(*b*) " Modulation," or the superimposition of the audio-frequency oscillations on a radio-frequency " carrier."

(*c*) " Detection," which is the process of abstracting the audio-frequency oscillations from the modulated carrier at the receiver.

(*d*) " Reproduction," which is the process of converting the audio-frequency electrical oscillations back into sound vibrations in the air.



(a) MODULATING CURRENT (A.F.)

(b) CARRIER WAVE (R.F.)

(c) MODULATED CARRIER WAVE (R.F.)

FIG. 6.

It will also be necessary to employ audio-frequency and radio-frequency amplification at intervals throughout the system, and to make use of the various methods of generating, radiating, and receiving the radio-frequency carrier, already explained for W/T transmission and reception.

It will be observed that operations (*a*) and (*d*) above are exactly opposite in principle, as also are (*b*) and (*c*). In line telephony, operations (*a*) and (*d*) alone are necessary.

These transformations are illustrated in Figs. 6 (*a*)-(*c*). Fig. 6 (*a*) shows the wave-form of the original and reproduced sound waves, and of the audio-frequency current. Fig. 6 (*b*) shows the carrier in an unmodulated state, and in Fig. 6 (*c*) the carrier is shown modulated by the A/F wave-form of Fig. 6 (*a*).

It should be noted that, after modulation and prior to detection, the audio-frequency currents do not exist as such; they are only represented by the variations in amplitude of the carrier. A **modulated carrier wave is a radio-frequency oscillation,** and corresponding circuits must therefore be employed for all operations after modulation and before detection.

The most common methods by which the above four operations may be effected will now be considered in more detail. The question of amplification at various stages is dealt with later.

9. **Types of Microphones.**—The microphone is an instrument for converting sound vibrations into corresponding audio-frequency electrical oscillations. It is nearly always followed by a valve amplifier, and the function of the microphone, and its immediately associated circuit, is to impress between the grid and filament of the first valve of the amplifier an oscillatory voltage whose wave-form corresponds to that of the original sound vibrations.

Until a few years ago the choice in microphones was limited to a few varieties of carbon microphone, being direct descendants from the first one of its kind invented by Hughes in 1878. With the advent of broadcasting and the talking film industry, these microphones proved not entirely satisfactory, and the result of consequent research is that the choice now includes :—

(*a*) The carbon microphone—the Post Office pattern and the transverse current type.
(*b*) The moving-coil microphone—including its converse the inductor type, and the ribbon microphone, as well as the ordinary moving coil.
(*c*) The condenser microphone.
(*d*) The piezo-electric microphone—also called the crystal microphone.

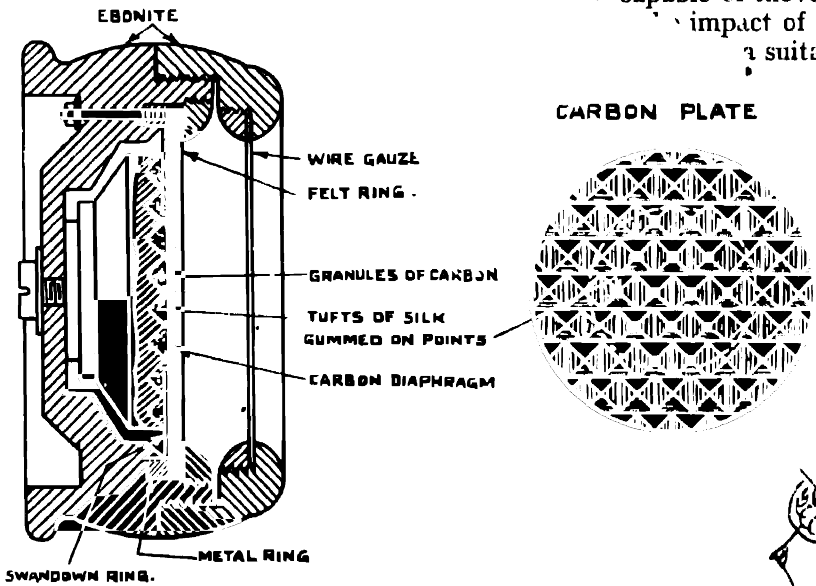
From the electrical point of view it may be seen that these microphones fall into three classes—electro-dynamic, piezo-electric, and capacity types. From the point of view of the directional behaviour of these microphones, it is also quite convenient and useful to classify them under two headings, namely, PRESSURE OPERATED types, and VELOCITY OPERATED types.

The latter classification will be understood if it is appreciated that sound waves travelling in air produce two effects at any point. As the rarefactions and compressions follow each other at a given point in air, so the air pressure at that point will oscillate about a normal value in accordance with the nature of the sound wave; these variations of pressure will take place at a point independently of the direction in which the sound is travelling.

Furthermore, since sound waves are longitudinal in nature, they produce a displacement of the air particles along the direction of propagation; they oscillate backwards and forwards about a zero position, and there will be no component of their velocity in direction at right angles to the direction of wave propagation. This implies that a microphone which depends for its action on the velocity of the air particles at a given point, will be highly directional in nature, responding only to the component of the velocity which lies along the direction of maximum sensitivity of the microphone; the theoretical horizontal polar response curve should take the form of a figure-of-eight diagram, as in the case of a rotating frame aerial (T.7).

Each type of microphone has its own field of usefulness, and a brief individual treatment is given below.

10. **The Carbon Microphone.**—A Service form of it is illustr: the density of packing of the diaphragm, or a rubber diaphragm coated on one side with carbon ıch the microphone is going to so that it lies closely to a carbon disc. These form the two electrodır one particular sound ir.tensity. being taken from them to the microphone terminals. The spaıear any strong magnetic field. If with carbon granules, which, under normal steady conditions, winplifier, screened and earthed cable sound waves impinge on the diaphragm, they cause it to vibrate ı on the carbon granules. It is a property of carbon that the resis varies approximately in an inverse ratio to the pressure between tlike small moving coil loudspeakers of the microphone will be varied in accordance with the impinginggrammatic representation, showing capable of movement across an impact of sound waves on a suitable amplifier.



Carbon Microphone.

FIG. 7.

These variations in microphone resistance have to produce corresponding oscillatory voltages between the grid and filament of the first valve of the amplifier, and so it is necessary to employ a constant external E.M.F., and a resistance or inductance in series with the microphone, as shown in Figs. 8 (*a*) and 8 (*b*). The inductance may be replaced by the primary winding of a microphone transformer, as shown in Fig. 8 (*c*), in order to take advantage of the voltage step-up provided by transformer coupling. The mathematical analysis of these circuits is here omitted. The practical considerations affecting them are :—

(*a*) The resistance R_s should normally have about two or three times the value of the resistance of the microphone.

(*b*) The impedance of the choke (L) in Fig. 8 (*b*), or of the primary of the transformer (T) in Fig. 8 (*c*), at the lowest frequency which it is required to reproduce efficiently, must be greater than the resistance of the microphone.

(*c*) For maximum sensitivity, the voltage of the battery (B) should be as large as possible, this being limited by the maximum current which the microphone can be allowed to carry without overheating and " packing " of the carbon granules.

The transformer method is much the most popular, since the oscillatory voltages developed across the secondary winding may be up to fifty times as great as those developed across a choke or resistance. There is no necessity, nowadays, for a microphone transformer to introduce any appreciable distortion. The use of a transformer also removes the necessity for a grid insulating

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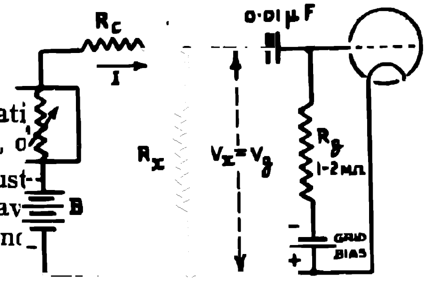
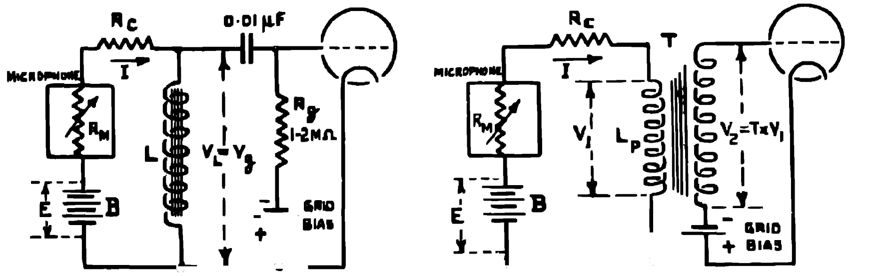


CIRCUIT EMPLOYING MICROPHONE RESISTANCE.

(a)



CIRCUIT EMPLOYING MICROPHONE CHOKE

(*b*)

CIRCUIT EMPLOYING MICROPHONE TRANSFORMER

(c)

CIRCUITS FOR USE WITH CARBON MICROPHONE.

FIG. 8.

condenser (Figs. 8 (*a*) and 8 (*b*)) and the consequent grid leak. These must be inserted when a choke or resistance is used, since the grid bias required on the first valve of the amplifier is usually quite different from the steady voltage developed across the resistance or choke.

The above variety of carbon microphone is essentially the same in operation as the Post Office pattern. In the case of the latter, the diaphragm is not in immediate contact with the carbon, but is connected at the centre, by a spindle, to a movable carbon electrode which varies the pressure on the carbon granules in the ordinary way. The spindle attachment of the diaphragm is frequently called a " button," and carbon microphones of the button type are very common.

In the " transverse current " type the diaphragm is made of non-conducting material; it is in contact with the carbon granules and conveys the alterations in air pressure directly to them. The current flows between two electrodes in the bed of carbon granules in a direction which is across the face of the diaphragm.

Carbon microphones are therefore " pressure operated " and essentially non-directional in nature. The best of the modern forms have extremely good frequency response characteristics, are very sensitive, and make satisfactory general purpose instruments if high quality of reproduction is not required, and the presence of a " background hiss " is not considered too objectionable. They are light and robust, and excellent for " hand sets."

The carbon microphone has the disadvantages that it will usually only work in one particular position (vertical, or nearly so), is very sensitive to external mechanical vibration, and requires a polarising current which must usually be obtained from a battery of about 4 to 10 volts. Also, the granules are liable to " pack," which necessitates turning the microphone upside down and

tapping it to unpack the granules. A further disadvantage is that the density of packing of the granules should depend on the average sound intensity under which the microphone is going to be used, and a microphone of this type is therefore only correct for one particular sound intensity.

If a microphone transformer is used, it must not be fitted near any strong magnetic field. If it is necessary to use a microphone at a long distance from the amplifier, screened and earthed cable should be used for the microphone lead.

11. **Moving Coil Microphones.**—In appearance these are like small moving coil loudspeakers and their principle of operation is the same. Fig. 9 (*a*) gives a diagrammatic representation, showing a small and light coil attached to a diaphragm, the whole being capable of movement across an intense radial magnetic field. As the " speech coil " moves, due to the impact of sound waves on the diaphragm, so variable E M F 's are generated in it, and are passed on to a suitable amplifier.

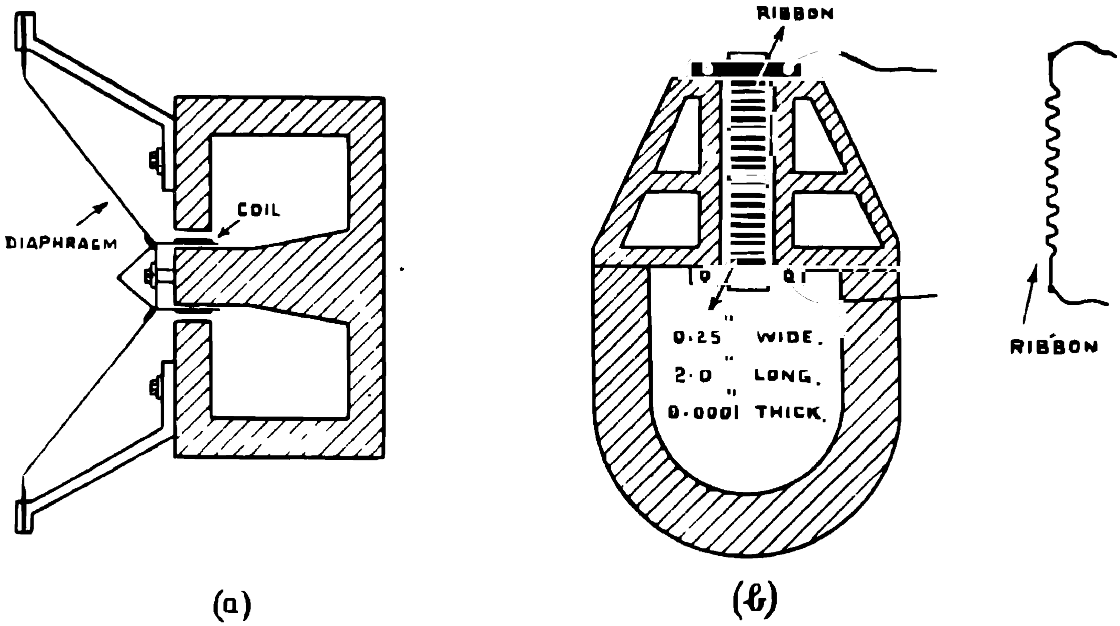


FIG. 9.

A form of the ordinary moving coil instrument is provided by the " inductor type." In that case the moving conductor, in which the small E.M.F. is generated, consists of a single tiny rod of aluminium rigidly coupled to a thin aluminium diaphram.

A further variety of the moving coil principle is seen in the more modern " ribbon microphone," represented diagrammatically in Fig. 9 (*b*). A very light ribbon, usually made of aluminium, is suspended between the poles of a magnet, electrical connection being made to its ends. It is usually folded like a concertina—or corrugated—in a direction at right angles to its length, in order to make the ribbon sufficiently flexible to allow free movement in a backwards and forwards direction, and to prevent any movement sideways in the direction of the pole pieces. The action of the ribbon may be compared with that of a single strand of wire moved at right angles to the direction of the magnetic flux ; whenever any movement occurs, an E.M.F. is induced in it which will, of course, be much smaller in amount than in the case of a moving coil.

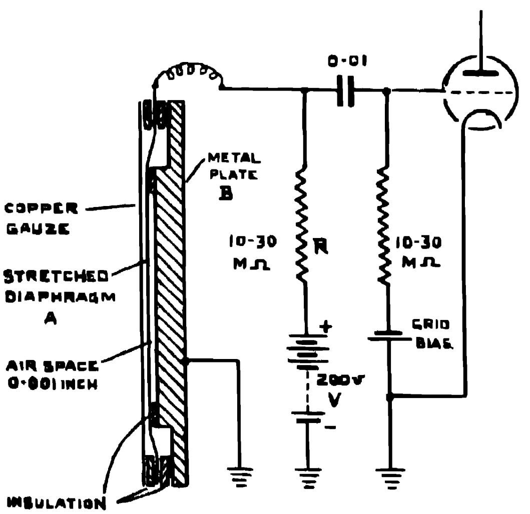
Ribbon type microphones have better response curves than the simple moving coil type ; this applies specially to the upper audio-frequencies. Their disadvantage is that considerable amplification is required of the small E.M.F which they generate.

With some ribbon microphones the ribbon is open to the air, both back and front; those will therefore be " velocity microphones," and possess figure-of-eight directional characteristics. Other types of ribbon microphone have only one side accessible to the sound waves, the other being terminated by an " acoustic impedance," which permits free movement of the ribbon by preventing reflection (*cf.* surge impedance). The effect of the latter is to produce a " pressure type " of ribbon microphone, having essentially circular directional properties in the horizontal plane.

Both pressure and velocity types of microphones have their uses in connection with talking films and broadcasting work. The B.B.C. has made considerable use of the velocity type for studio work, but the talking film industry usually finds that a pressure type microphone is simpler in operation.

The advantages of the simpler type of moving coil microphone are—robustness, constancy of performance, freedom from " packing," cannot be overloaded by shouting, relatively insensitive to external mechanical vibration, can be held in any position, and does not require any polarising current. Except in the case of the low impedance ribbon type of instrument, it can be connected to its transformer by a twin core cable which can be almost any length, but which should be screened and earthed if more than 100 ft. or so in length, to avoid pick-up of extraneous noises. The transformer ratio must be arranged to match the impedance of the speech coil—about 25 ohms in low impedance models—with the input impedance of the grid/filament circuit of the first valve of the amplifier. It is important that the connections between the secondary of the transformer and the grid/filament circuit should be as short as possible.

12. **The Condenser Microphone.**— A typical condenser microphone is shown in Fig. 10. It consists, essentially, of a thin tightly-stretched diaphragm (A) of aluminium alloy, secured around its edge so that it lies very close to, but well insulated from, a solid metal plate (B). These form the electrodes of the microphone. A steady D.C. polarising P.D. (V) of about 200 volts, is applied between the electrodes through a high resistance R, which may have a value of 20 megohms or more. When sound waves impinge on the diaphragm (A), they cause it to vibrate, and so to vary the capacity between the electrodes. The resistance R is sufficiently large to prevent any change in the quantity of electricity (Q) stored in the capacity of the microphone, except at a very slow rate. Consequently, it will be seen from the definition of capacity, $C = \frac{Q}{V}$, that since Q is constant, any changes in the capacity (C), such as are caused by impinging sound waves, will cause a corresponding change in V. These voltage variations will be equal and opposite across the microphone and the resistance, and the grid-filament circuit of the first valve of the microphone amplifier may therefore be connected across either.



CONDENSER MICROPHONE & ASSOCIATED CIRCUIT.

FIG. 10.

In historical development, the condenser microphone succeeded the carbon granule type and was much used for broadcasting work. It possesses an extremely uniform response over the major portion of the audio-frequency band, and is entirely free from " background hiss." It is, however, comparatively insensitive, and in comparison with the carbon microphone, the condenser microphone requires about two extra stages of amplification. No impedance matching transformer is needed.

13. The Piezo-Electric Microphone.—This is a comparatively new product, its principle of action depending upon the use of crystals having piezo-electric properties, of which Quartz and Rochelle salt are two examples (K.42). In this case, Rochelle salt is commonly used since it is many times more sensitive than Quartz. There are two types of microphone employing this principle, the " diaphragm " type, and the " sound cell " type. (There are also loudspeakers and gramophone pick-ups.)

The diaphragm type is the simpler and less expensive variety. Two square crystal slices are cemented together, back to back, in such a way that if a voltage is applied between the crystal faces, the effect will be to make one slice tend to expand and the other to contract. This would produce an effect analagous to the bending of a beam supported at the two ends with a weight placed in the middle. In this case, the converse action takes place. An ordinary diaphragm with a button and a spindle conveys the variations in acoustic pressure to the centre of the crystal slice, which is supported so that it can be bent up and down like a beam; as it bends so it generates piezo-electric voltages which can be passed on to an amplifier in the ordinary way. This type of microphone gives an output equal to the best carbon microphones, and, in addition :—

(*a*) The frequency response curve is extremely good and is particularly suitable for high quality reproduction of speech.

(*b*) The microphone can be held in any position.

(*c*) It does not suffer from the phenomenon of " packing " (*cf.* carbon microphone).

The piezo-electric microphone is not well adapted for use in hot climates, since the crystals tend to deteriorate at temperatures above about 120° F.

The sound cell type is an adaptation of the above simple type, produced in order to increase the output and make the response more perfect. Fig. 11 gives a diagrammatic representation of the essential features of a single sound cell unit; it is a vertical section of a unit which is square in shape, with sides about 1 inch in length. The crystal plates are about 0·5 inch square and 0·2 inch thick, the resonant frequency being over 12,000 cycles per second. Two plates are cemented together—as above—in order to form a unit. Two of these units are then mounted in a rectangular frame of insulating material, an enclosed air space being left between the crystal units in order to provide freedom of movement. Finally, the space containing the crystal units is made airtight and moisture proof. Variations in sound pressure act on suitable diaphragms which transmit the variations to the crystal units; the voltages developed across the crystal faces of the two units are arranged to be in phase.

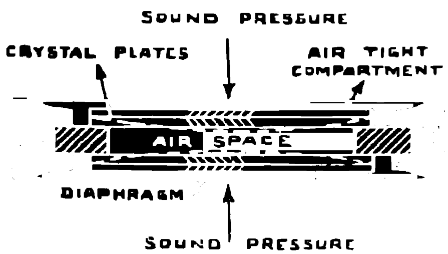


FIG. 11.

In practice, a number of these small square sound cells may be arranged vertically one above the other, and their outputs may be connected in series or in parallel or a combination of the two. The whole sound cell assembly is usually surrounded by a grille.

Since these sound cell microphones are pressure operated, they have polar response curves which are non-directional in character. Their impedance is high and they can be directly connected to the grid of the first valve of an amplifier, employing a grid leak of about 5 megohms.

14. Modulation : Terminology.—Modulation is the process whereby the audio frequency electrical oscillations produced by the microphone circuit are superimposed on the radio frequency C.W. oscillation called the carrier; the resultant waveform is known as a " modulated wave." The audio frequencies no longer exist physically; their intensity at any instant is represented by the difference between the actual amplitude of the carrier at that instant and its unmodulated amplitude. A modulated wave is entirely a radio frequency oscillation, and prior to detection it exhibits no physical audio frequency properties. This has already been illustrated in Fig. 6

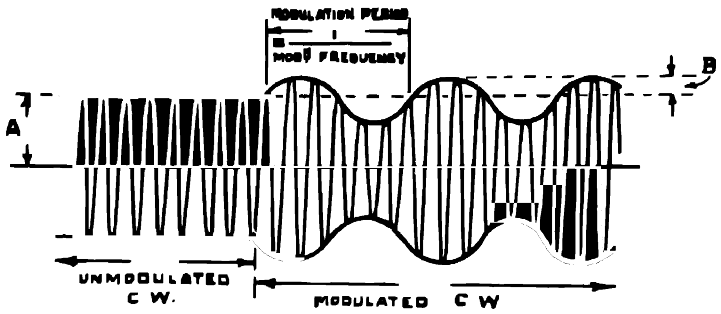


FIG. 12.

Fig. 12 shows a carrier wave of unmodulated amplitude A, modulated by a pure tone, *i.e.*, the modulated wave is of the type called Tonic Train. The variation of the carrier amplitude is sinoidal, and takes place between the limits A + B and A — B. B is called the "Depth of Modulation."

The ratio $\frac{B}{A}$ gives a measure of the degree of modulation applied to the carrier amplitude. Expressed as a percentage, *i.e.*, $\frac{100B}{A}$, it is known as the "Percentage Modulation" (N).

For 100 per cent. modulation, (B = A), the amplitude of the modulated carrier varies from zero to double its unmodulated amplitude. In R/T, the percentage modulation should never be allowed to exceed about 80 per cent., for reasons to be discussed below.

15. **Sidebands.**—Instead of regarding the modulated carrier wave as an R/F oscillation of constant frequency and varying amplitude, it is often more convenient to consider it as the resultant of a number of superimposed R/F oscillations which differ both in amplitude and frequency from each other, but each of which is of constant amplitude and frequency, *i.e.*, possesses a C.W. waveform. The sinoidally modulated carrier wave of Fig. 12, for instance, may be considered as the resultant of three C.W. oscillations whose frequencies are the frequency of the carrier, and the frequency of the carrier plus and minus the modulating frequency respectively.

This result may be derived as follows. The unmodulated carrier, of frequency $\frac{\omega}{2\pi}$, and amplitude A, is represented by the expression A sin ωt. With a modulating frequency $\frac{p}{2\pi}$, and a depth of modulation B, the amplitude of the modulated carrier (the envelope of the oscillation of Fig. 12) may be expressed as A + B sin pt, and so the complete expression for the modulated carrier becomes (A + B sin pt) sin ωt.

On expansion this becomes A sin ωt + B sin pt sin ωt

$$= A \sin \omega t + \frac{B}{2} \cos (\omega - p)\, t - \frac{B}{2} \cos (\omega + p)\, t$$

$$= A \sin \omega t + \frac{B}{2} \sin \left[(\omega - p)\, t + \frac{\pi}{2} \right] + \frac{B}{2} \sin \left[(\omega + p)\, t - \frac{\pi}{2} \right] \quad \ldots\ldots\ldots\ldots \quad (1)$$

The modulated carrier is thus equivalent to three C.W. oscillations :—

(1) A sin ωt, whose amplitude and frequency are those of the unmodulated carrier.

(2) $\frac{B}{2} \sin \left[(\omega - p)\, t + \frac{\pi}{2} \right]$, whose frequency is less by the modulation frequency than that of

the unmodulated carrier, and whose amplitude is half the depth of modulation. It is known as a " Lower Sideband " oscillation.

(3) $\frac{B}{2} \sin \left[(\omega + p) t - \frac{\pi}{2} \right]$, whose frequency is **greater** by the modulation frequency than that of the unmodulated carrier, and whose amplitude is half the depth of modulation. It is known as an " **Upper Sideband** " oscillation.

It is to be observed that all three components are R/F oscillations.

It has been seen that the complex vibrations of speech and music can be resolved into a number of pure tones similar to that represented by B sin pt. Hence, when a carrier wave is modulated by the audio frequency electrical oscillations corresponding to these complex vibrations, the modulation of the carrier by each component pure tone is equivalent to the production of three C.W. radio frequency oscillations. The wave radiated may therefore be analysed into a C.W. oscillation at the carrier frequency, and a large number of C.W. oscillations whose frequencies are greater and less than that of the carrier by amounts equal to the frequencies of the original component pure tones in the speech or music. The band of such frequencies above the carrier frequency is known as the " Upper Sideband," and that below the carrier frequency as the " Lower Sideband," *e.g.*, if the carrier frequency is 1,000 kc./s. and the limiting frequencies of the speech or music are 50 and 8,000 cycles per second, the frequency range of the upper sideband is from 1,000,050 to 1,008,000 cycles per second, and that of the lower sideband is from 999,950 to 992,000 cycles per second.

A modulated wave thus covers a considerable band of frequencies, the width of the band being twice the highest modulation frequency (16,000 cycles per second in the example above). A receiving circuit for such waves must therefore be adjusted so that all frequencies in the band are equally well received, and to prevent interference in reception the difference between the carrier frequencies of any two R/T transmitters must be more than the sum of their highest modulation frequencies. The greater the number of R/T transmitters that have to be accommodated in a given range of frequencies, the less is the highest modulation frequency that can be allowed if interference is to be avoided. In commercial broadcasting, the highest normal modulation frequency is 8,000 cycles per second. The difference between carrier frequencies should then be about 20 kc./s. for stations likely to interfere with each other. This is not at present the case in practice.

In January, 1934, the provisions of the Lucerne Conference came into force. This arranged for an interval of 9 kc./s. between the carrier waves corresponding to adjacent broadcasting stations, care being taken that those stations which are adjacent in frequency are widely spaced geographically. The scheme was a compromise, for without wide geographical separation it is quite impossible for two high powered R/T stations to operate without causing mutual interference, unless the highest modulating frequency were to be restricted to 4,500 cycles per second.

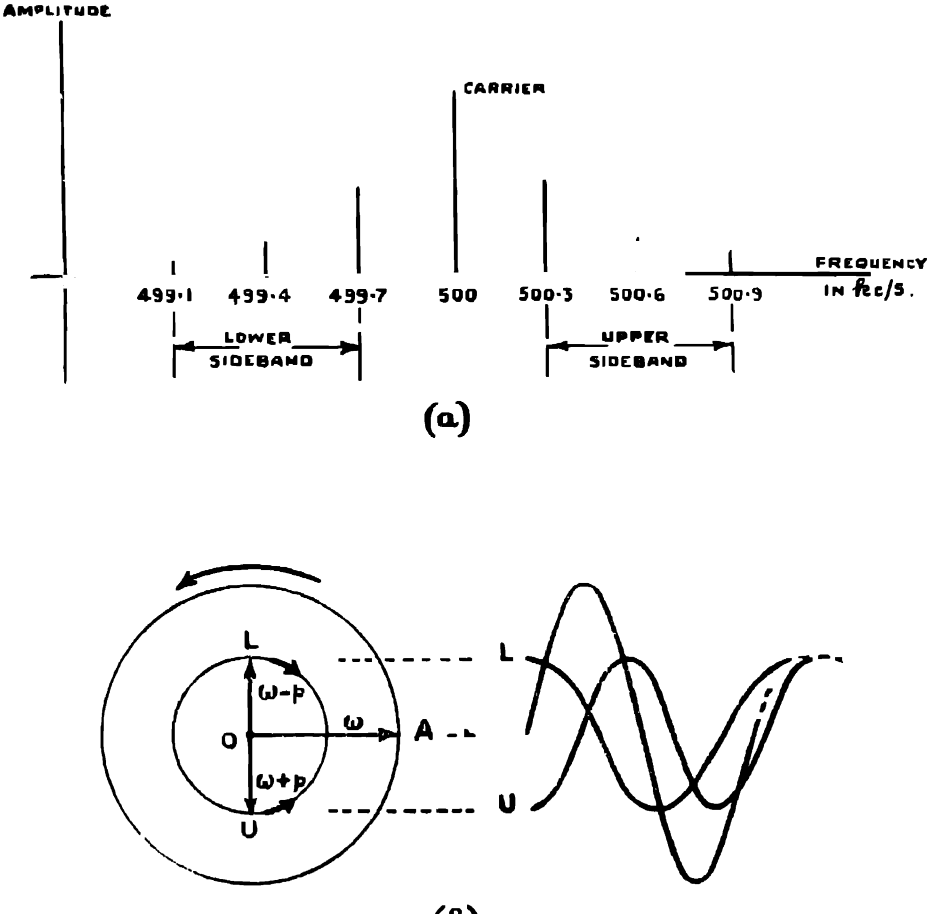
The allocation of channels 9 kc./s. in width, in the case of each R/T station, may be contrasted with the allocation of channels 6 Mc./s. in width for the use of television transmitters. The latter is due to the much greater width of the sideband frequencies, extending on the average 2·5 Mc./s. above and below the carrier frequency. Moreover, since it is essential to have suitable frequency gaps between the carrier used for the transmission of video-frequencies and that used for the transmission of the corresponding audio-frequencies, the practical width of the frequency band occupied by one station amounts to about 6 Mc./s. For example, in the case of the Alexandra Palace transmitter, the band width extends from 41·5 Mc./s.—the carrier used for SOUND—to 47·5 Mc./s.—the limit of the upper sideband corresponding to the highest video-frequency modulation ; 45 Mc./s. is the frequency of the carrier used for VISION. Commercial television receivers are designed to have characteristics which pass a band of frequencies 5 Mc./s. in width at the carrier frequency.

The upper and lower sideband frequencies may also beat with each other and produce still higher frequencies ; this is undesirable on account of the greater tendency to interference and the distortion introduced. This effect, however, can be made negligible by limiting the percentage modulation to about 80 per cent.

The sidebands produced when a carrier wave at 500 kc./s. is modulated by the wave form of Fig. 2, are shown in Fig. 13. The first harmonic of the violin note is taken as 300 cycles per second.

It will be seen that each component of the modulating frequency produces its own sideband oscillation, and that the amplitudes of the sideband oscillations are proportional to the amplitudes of the component frequencies producing them.

Thus, in Fig. 13, the sideband amplitude of the fundamental modulating frequency is greater than the sideband amplitudes of the weaker harmonics.



(b)

FIG. 13.

The relation between sidebands and carrier may also be demonstrated by means of curves and generating vectors, as in Fig. 13 (*b*). With reference to equation (1), it will be seen that at time $t = 0$, the generating vector L for the lower sideband may be drawn leading that for the carrier by 90°, the upper sideband vector U lagging by the same amount. The *relative* angular velocity of each sideband vector with respect to the carrier vector is given by $\omega - (\omega - p)$ and $\omega - (\omega + p)$, *i.e.*, p and $-p$. With respect to the carrier vector, this is equivalent to a *relative* clockwise rotation of the lower sideband vector, and a *relative* anti-clockwise rotation of the upper sideband vector. The whole system of vectors is to be regarded as rotating in the usual anti-clockwise direction, an effect which may be pictured by regarding the three vectors as being drawn on a piece of cardboard having the initial phase relationships shown in the sketch, the whole being

rotated about an axis through O. Considering the relative velocities, since each sideband vector moves through the same relative angle in the same time, it is evident that they will be simultaneously in phase with the carrier, and simultaneously out of phase with the carrier half a period later. The sum of the three curves, the initial portions of which are given in Fig. 13 (*b*), will produce the familiar modulated wave form. The maximum and minimum amplitudes occur when the sideboard vectors are in phase or in anti-phase with the carrier.

This vector concept may be used to account for one of the difficulties which is now being experienced in connection with television. Short cable links between the studio and the television transmitter are unavoidable, and considerable difficulty is experienced in providing a cable link in which there is no undue attenuation of the very high frequencies in use. Moreover, it has been recently shown in Germany that *as the frequency is increased*, the skin effect becomes more pronounced and *the velocity of the guided wave increases*. It has been shown that there is an appreciable difference in velocity between wave frequencies at opposite ends of the sideband spectrum at television frequencies. With reference to Fig. 13 (*b*), the effect of this difference in velocity would be that at *the generator end* of a cable, the two sideband vectors would coalesce at the moment when they are in phase or in anti-phase with the carrier, but that at *the load end* the resultant of the two vectors would be no longer in line with the carrier vector. As the length of the cable link increases the upper sideband vector continues to gain in phase, the lower sideband vector losing in phase in a similar manner. There will be a critical length of cable at which the upper sideband has gained 90° in phase; the sideband vector resultant is then at right angles to the carrier vector and the modulation of the carrier tends to disappear. This is one of the factors which sets a limit to the distance over which television frequencies may be transmitted by cable.

16. **Power in the Sidebands.**—With reference to the work of paragraph 15, it will be apparent that the total energy in a modulated wave is the sum of the energy contents of the various components.

Consider the simple case of a carrier A sin ωt modulated by a single audio-frequency B sin pt. If the modulation is 100 per cent., we have A = B, and the three components A sin ωt, $A/2 \sin\left[(\omega - p)\,t + \frac{\pi}{2}\right]$, $A/2 \sin\left[(\omega + p)\,t - \frac{\pi}{2}\right]$, are each radiated.

An estimate of the relative power expended in these three components is obtained by regarding A (above) as the amplitude of the aerial current ($\mathcal{I}$).

Hence Power $\propto \mathcal{I}^2 \propto A^2$. This gives

Power in carrier $\propto A^2$

Power in lower sideband $\propto \left(\frac{A}{2}\right)^2 = \frac{A^2}{4}$

Power in upper sideband $\propto \left(\frac{A}{2}\right) = \frac{A^2}{4}$

Total power in sidebands $\propto \frac{A^2}{2}$ .... and we may write

Total power radiated $\propto A^2 + \frac{A^2}{2}$ = carrier power $(1 + \frac{1}{2})$.

From this it appears that, of the total power radiated, two-thirds of it is contained in the carrier and one-third in the side bands. Practical proof of the above is given by the fact that the carrier wave of a distant R/T station may often be received virtually bereft of its sidebands; assuming equal attenuation of carrier and sidebands, the latter will be the first to disappear below the noise level of the aerial and the receiver (*cf.* Signal/noise ratio).

Considering a numerical example :—A transmitter supplies 10 kW. to the aerial when unmodulated; what will be its output when modulated 100 per cent.? Since the power in the carrier

remains unchanged, 15 kW. must be the result, if one-third of the total power is expended on the sidebands.

In the more general case, where the percentage modulation (N) is less than 100 per cent., the total power in the sidebands will be given by $\frac{B^2}{2}$ or $\frac{(NA)^2}{2}$. A formula can then be written—

$$\text{Total power radiated} \propto A^2 + \frac{B^2}{2} = A^2 + \frac{(NA)^2}{2}$$

$$= \text{Carrier power}\left(1 + \frac{N^2}{2}\right) \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

e.g., with 80 per cent. modulation of a 10 kW. carrier—

$$\text{Total power radiated} = 10\left(1 + \frac{0{\cdot}8^2}{2}\right) = 13{\cdot}2 \text{ kW.}$$

For this reason the percentage modulation should be kept as high as possible (about 80 per cent.).

It should be realised that all the "intelligence," whether it be speech or music, is conveyed by the sidebands, the carrier playing an essential but negative part from this point of view. In general, the percentage modulation will vary as the intensity of speech or music varies about its mean value, the excess power in the carrier serving as a reservoir of energy to supply the demands of the signal for sideband energy.

The above work may be extended to include the case of the single sideband suppressed carrier system (S.S.B. system) of telephony (paragraph 17). For the purposes of comparison, it will be assumed that the amplitude of the aerial current corresponding to the single sideband is adjusted to give the SAME ACOUSTIC SIGNAL STRENGTH at the receiver, as that produced by the double sideband system (D.S.B. system), the amplitude of the "injected carrier" being similarly equivalent; in practice, the latter is always greater.

From the work on "square law detection" in Section "D," it will be recalled that the rectified anode current varies as the SQUARE OF THE SIGNAL VOLTAGE INPUT; in the case of heterodyne action there are two signal inputs, and the A/F change in anode current depends on the PRODUCT OF THE SIGNAL VOLTAGE INPUTS.

Let D be a constant of detection, depending on the detector.

D.S.B. system.

Field strength amplitude due to each sideband $\propto \frac{NA}{2}$.

When each sideband beats with the carrier of amplitude A, an audible signal results, and

$$\text{Acoustic signal strength} = 2D\,\frac{NA}{2} \times A = DNA^2. \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (2)$$

Since the MEAN carrier power varies as $\frac{A^2}{2}$ .... (Vol. I).

From (1) .... MEAN power radiated $\propto \frac{A^2}{2}\left(1 + \frac{N^2}{2}\right)$ .... (3)

or, ... Mean power in each sideband $\propto \frac{A^2N^2}{8}$ .... (4)

S.S.B. system.

Let B_1 be the field strength amplitude of the received sideband; "injecting" a carrier of amplitude A we have—

$$\text{Acoustic signal strength} = DAB_1. \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (5)$$

Now the acoustic signal strengths (2) and (5) will be equal when $B_1 = NA$. When this is the case, the S.S.B. amplitude $= 2 \times$ amplitude of either sideband in the D.S.B. case; with DOUBLE the field strength amplitude, the transmitted power must be FOUR TIMES greater than that in each sideband of the D.S.B. case—

$\therefore$ From (4) Mean power radiated $\propto \frac{A^2N^2}{2}$.

Hence

$$\frac{\text{Power in S.S.B. system}}{\text{Power in D.S.B. system}} = \frac{A^2N^2/2}{\frac{A^2}{2}\left(1 + \frac{N^2}{2}\right)} = \frac{2N^2}{2 + N^2}.$$

For $N = 100$ per cent., the above ratio is two-thirds and represents a power saving of one-third. This ratio is unchanged if linear detection is substituted for square law.

In practice, the mean value of N during speech is about 20 per cent., but peaks occur which are much higher. It should also be noted that the necessity for preserving an equivalent signal/noise ratio (P.15) modifies the above reasoning and it can be shown that the above ratio is more nearly given by the formula $\frac{N^2}{4 + 2N^2}$.

17. **Modulation: Alternative Methods of Transmitting Intelligence.**—It has been shown that the effect of modulating a carrier wave of frequency F by means of an audio-frequency N is to produce a result equivalent to three simple oscillations of frequency F, F + N, and F — N. It will also be appreciated that although the carrier is essential to the production of the sidebands, the receipt of the carrier alone will not convey any intelligence to the listener; a rectified C.W. signal—such as the carrier— simply increases the steady value of the mean anode current and no sound is heard.

The audio-frequency sounds which constitute the "intelligence" are entirely contained in the sidebands, and it should be noted that the A/F modulation is the beat or difference frequency between any sideband oscillation and the carrier wave. It is this beat frequency between the carrier and its sidebands which appears after detection; the operation of modulating a carrier wave may be considered equivalent to the displacement of a band of audio-frequencies from their normal position in the frequency spectrum to a corresponding position at a higher frequency at which energy may be satisfactorily radiated by an aerial.

From the foregoing remarks the suggestion is irresistible that, in order to produce the requisite beat frequencies after detection, we only need to have the carrier wave and one of its sidebands. It would appear, moreover, that even the carrier need not be radiated, since it could fairly easily be supplied at the receiver by means of a local oscillator. There appear to be the following alternative methods of transmitting intelligence by modulating a carrier wave :—

(*a*) Transmission and reception of both sidebands and the carrier. This is the present practice in normal broadcast R/T.

(*b*) The transmission and reception of one sideband only, the correct carrier being replaced at the receiver. This method has been used in some commercial R/T work.

(*c*) The transmission and reception of one sideband and the carrier. This system possesses the advantages of that in (*b*) and, in addition, it avoids the somewhat difficult operation of replacing the carrier at the receiver.

(*d*) The transmission and reception of both sidebands, the carrier being replaced at the receiver. Of the various alternatives, this is the one most fraught with difficulty.

Methods (*b*) and (*c*) are "single sideband" systems, the adoption of which afford certain advantages and disadvantages. From the work of paragraph 16, it is clear that two-thirds of the power of a 100 per cent. modulated wave is contained in the carrier, the radiation of which adds nothing to the intelligence received. If one sideband is transmitted, having sufficient power to

give an acoustic signal equal to that given by a carrier and both sidebands, it has been shown (paragraph 16) that the single sideband transmitter will consume about two-thirds of the power dissipated by a similar transmitter radiating both sidebands and the carrier, making the assumption of 100 per cent. modulation. Another obvious advantage of a single sideband system is the decrease in width of the requisite frequency band to half of that needed by the double sideband system. This gives room for further channels of communication in the frequency spectrum. Moreover, the various H/F circuits may be more sharply tuned, and the aerial efficiency may therefore be increased.

From the ordinary broadcasting point of view, the replacement of the carrier at the receiver constitutes a difficulty. In general, it must be correctly replaced with an error of only $\pm$ a few cycles per second This is a very severe requirement for the simple broadcast receiver, but it certainly constitutes a possible development. The radiation of the carrier with one sideband represents a way of overcoming the above difficulty It would be distinctly easier from the point of view of the simple receiver, while at the transmitter no insuperable problem would be involved.

It is proposed to consider briefly some of the problems involved in single sideband telephony.

In addition to the foregoing methods of transmitting intelligence by **amplitude modulation,** there is the possibility of doing it by keeping the amplitude of the wave constant and varying the frequency in accordance with the signal to be sent. This is called **frequency modulation,** reception of the frequency modulated wave being effected by conversion into amplitude modulation The tuned circuits of the receiver must be deliberately de-tuned so that the signal frequency is slightly away from the peak of the response curve ; the response of the resonant circuits varies with the frequency of the incoming signal, so producing ordinary amplitude modulation from the incoming frequency modulated wave. This system is at present little used ; it appears to have few advantages, and its sideband spectrum is at least as large as that produced by amplitude modulation.

18. **Single Sideband Telephony—The Balanced Modulator.**—The foregoing work has pointed out the possibility of saving power at the transmitter, and certain other advantages to be obtained by the adoption of a single sideband system. In this connection it should be noted that the power saving is due to the suppression of the carrier, the suppression of one of the sidebands contributing nothing in this respect

Suppression of the carrier frequency may be achieved in various ways, one of which involves the use of a "balanced modulator," represented diagrammatically in Fig. 14. The action of the circuit resembles the heterodyne detector, or beat rectifier circuit, considered in the D/F section (T.10).

BALANCED MODULATOR

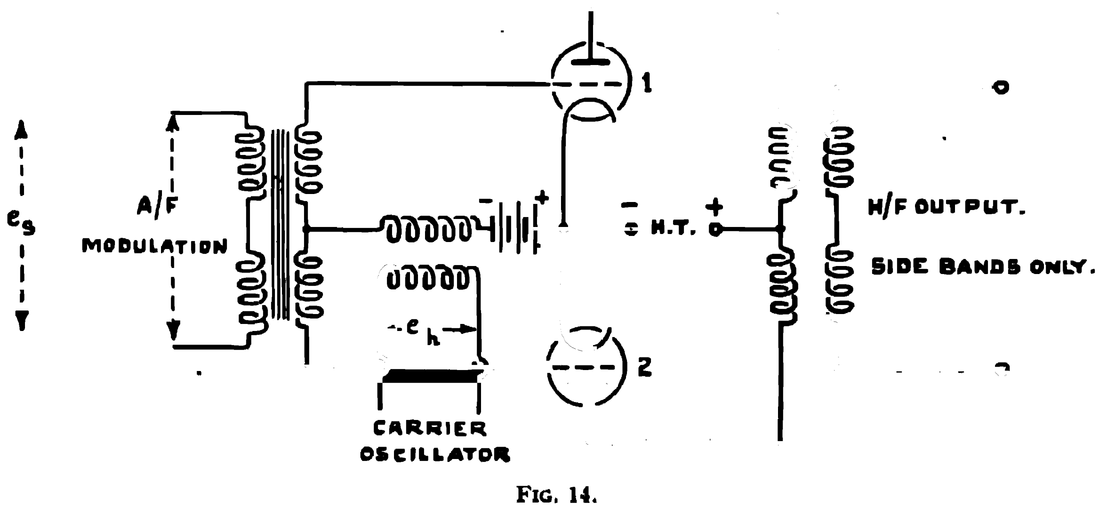


FIG. 14.

The push-pull arrangement of valves has two inputs, an ANTI-PHASE INPUT (e_s) constituted by the A/F modulation, and a PARALLEL INPUT (e_k), due to the carrier frequency; quite clearly, the parallel input operates the two grids in phase with each other. The valves are operated as detectors, using " lower anode band detection " in this case. On the output side, two identical anode coils are linked magnetically to a third circuit.

When the circuit is operating, the total input to each valve is the sum of the oscillatory parts and the steady pre-signal voltage. If the parallel input is instantaneously in phase with the anti-phase input at the grid of valve 1, it will, at the same time, be in anti-phase with it at the grid of valve 2. The effect of these inputs to the push-pull combination, with the valves operating as detectors, is to produce R/F components in the output circuit having frequencies corresponding to the sum and the difference of the carrier frequency and the modulating frequency, together with a number of other components. The presence of sum and difference components passed on with the carrier frequency, indicates the production of the sideband oscillations which are normally the result of any modulating process. Modulation has, in fact, been achieved, the carrier frequency and its sidebands both being present in the output circuit. The anode coils are linked magnetically to a third circuit and, since the two anode currents are relatively in opposite directions, the flux produced by each component of the one is in opposition to that due to each component of the other; the nett effect on the coupled circuit is that due to the difference of the flux produced by the current in each half of the output circuit. Mathematically, it is easy to show that, with perfect electrical balance, this involves the complete cancellation of the carrier frequency, the sideband oscillations alone being passed on to suitable amplifiers and filters. The possibility of this being the case is suggested if one considers that the carrier frequency is the only input to the system. In that case, it is clear that the oscillatory currents in the output circuit are in anti-phase, the flux produced by the current in one-half of the output impedance cancelling that produced by the current in the other. An abbreviated mathematical analysis is contained in paragraph 19.

The single sideband system is in use on the Long Wave Transatlantic Telephone Service operated from Rugby. In that case it is only desired to produce intelligible speech, for which reason it is only necessary to use modulating frequencies between 250 and 2,700 cycles per second. This has the effect of reducing the width of the sideband frequencies, the mechanism of the process being illustrated in Fig. 15. For simplicity it may be assumed that the modulating frequencies

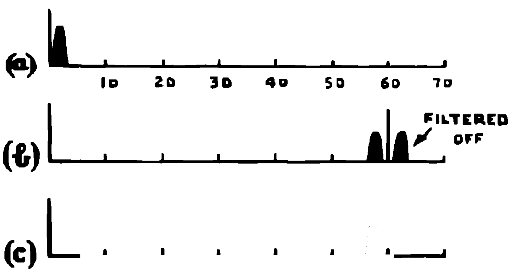


FIG. 15.

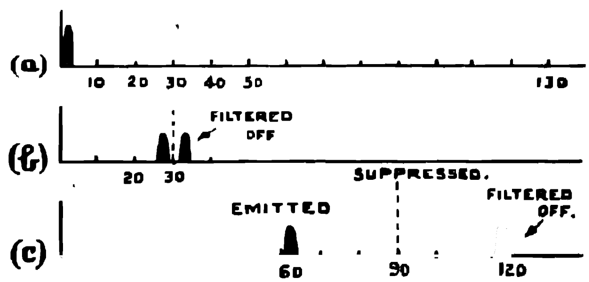


FIG. 16.

extend over the band 0-3,000 cycles per second, and that the carrier frequency is 60 kc./s. Using the balanced modulator a band of frequencies, covering the range 57-63 kc./s. (Fig. 15 (*b*)), will be passed on to suitable filters and amplifiers. The function of the filter is to eliminate frequencies corresponding to one of the sidebands, and, in this practical case, those corresponding to the upper sideband—60-63 kc./s.—are removed. The effect is shown diagrammatically in Fig. 15 (*c*).

As one might expect, the suppression of frequencies so closely adjacent cannot be perfectly achieved with any simple filter and, in practice, the production of oscillations corresponding to the lower sideband is effected in two steps, shown diagrammatically in Fig. 16. The speech frequency band is used to modulate a carrier frequency lower than the radiation frequency, using the balanced

modulator, a carrier frequency of about 30 kc./s. being used to illustrate this practical case. This produces sideband frequencies extending from 27 kc./s. to 33 kc./s., the upper sideband being filtered off considerably more easily than when the carrier frequency is higher. The lower sideband frequencies, 27 kc./s. to 30 kc./s. are now used to modulate a carrier (say), 90 kc./s., using the balanced modulator. This produces sideband oscillations covering the ranges 60-63 kc./s., and 117-120 kc./s. These two frequency bands are widely apart, and it is easy to design filters to effect their separation. The final result of this "double modulation" process is the radiation of a band of frequencies in the neighbourhood of 60 kc./s.

At the receiving end it is necessary to re-apply the carrier frequency which was suppressed at the transmitter, in order to produce beat notes with the sideband frequencies restoring them to their proper position in the frequency spectrum—Fig. 16 (*a*). The effect of re-applying the carrier is termed "de-modulation," and just as the modulation is done in two stages, so, in certain cases, the de-modulation may be effected in two steps. The necessity for the restoration of the carrier with considerable accuracy; adds some degree of secrecy to the system; for commercial speech an error of $\pm$ 20 cycles is permissible.

★19. **Mathematical Treatment of the Balanced Modulator.**—The anti-phase input (e_s) is applied equally to the two grids, and is here expressed by $\mathcal{V}_s \sin \omega_2 t$. The parallel input ($e_h$) is applied in the same phase between the grid and cathode of each valve in the balanced circuit, and is here expressed by $\mathcal{V}_h \sin \omega_1 t$.

Over a relatively short range any curve may be represented by a parabolic formula; the anode current could, therefore, be given by an equation of the form—

$$I_a = a + bV_g + cV_g^2,$$

where a, b and c are constants.

When there is no oscillatory input to the grids, the steady "pre-signal voltage" may be denoted by V_0, and the steady anode current in each valve will be given by—

$$I_0 = a + bV_0 + cV_0^2,$$

provided that the valves are identically the same.

When the circuit is operating, the total input to each valve is the sum of the oscillatory part and the steady pre-signal voltage; it is given by—

$V_0 + \mathcal{V}_h \sin \omega_1 t + \mathcal{V}_s \sin \omega_2 t$ .... between grid and cathode of (say), valve (1),

and

$V_0 + \mathcal{V}_h \sin \omega_1 t - \mathcal{V}_s \sin \omega_2 t$ .... between grid and cathode of valve (2).

The corresponding anode currents will be—

$$\begin{aligned}
i_1 &= a + b(V_0 + \mathcal{V}_h \sin \omega_1 t + \mathcal{V}_s \sin \omega_2 t) + c\,(V_0 + \mathcal{V}_h \sin \omega_1 t + \mathcal{V}_s \sin \omega_2 t)^2 \\
&= a + bV_0 + cV_0^2 + b\mathcal{V}_s \sin \omega_2 t + 2cV_0\mathcal{V}_s \sin \omega_2 t \\
&\qquad + 2c\mathcal{V}_h\mathcal{V}_s \sin \omega_1 t.\ \sin \omega_2 t + \text{other positive H/F terms} \\
&= I_0 + b\mathcal{V}_s \sin \omega_2 t + 2c\,V_0\mathcal{V}_s \sin \omega_2 t \\
&\qquad + 2c\mathcal{V}_h\mathcal{V}_s\,[\cos(\omega_1 - \omega_2)\,t - \cos(\omega_1 + \omega_2)\,t] \\
&\qquad + \text{other positive H/F terms} \quad \ldots\ldots \quad (1)
\end{aligned}$$

and similarly

$$\begin{aligned}
i_2 &= I_0 - b\mathcal{V}_s \sin \omega_2 t - 2cV_0\mathcal{V}_s \sin \omega_2 t \\
&\qquad - 2c\mathcal{V}_h\mathcal{V}_s\,[\cos(\omega_1 - \omega_2)\,t - \cos(\omega_1 + \omega_2)\,t] \\
&\qquad + \text{other positive H/F terms.} \quad \ldots\ldots \quad (2)
\end{aligned}$$

The anode coils are linked magnetically to a third circuit, and since the two anode currents are relatively in opposite directions, the flux produced by each component term of the one is in opposition to that due to each term of the other: the total effect is represented by taking the difference of (1) and (2). WITH A PERFECTLY BALANCED CIRCUIT, the terms which are positive in each expression will disappear, leaving—

$$i_1 - i_2 = 4c\mathcal{V}_s\mathcal{V}_h \cos(\omega_1 - \omega_2)\,t - 4c\mathcal{V}_s\mathcal{V}_h \cos(\omega_1 + \omega_2)\,t + (2b\mathcal{V}_s + 4cV_0\mathcal{V}_h) \sin \omega_2 t.$$

The first two terms in the expression above represent the sideband frequencies, one of which is filtered off. The third term is at audio-frequency, and is, therefore, not radiated.

In the Service use of this circuit in D/F work, the difference frequency component is passed into a circuit of fixed tuning and thence to an intermediate frequency amplifier.

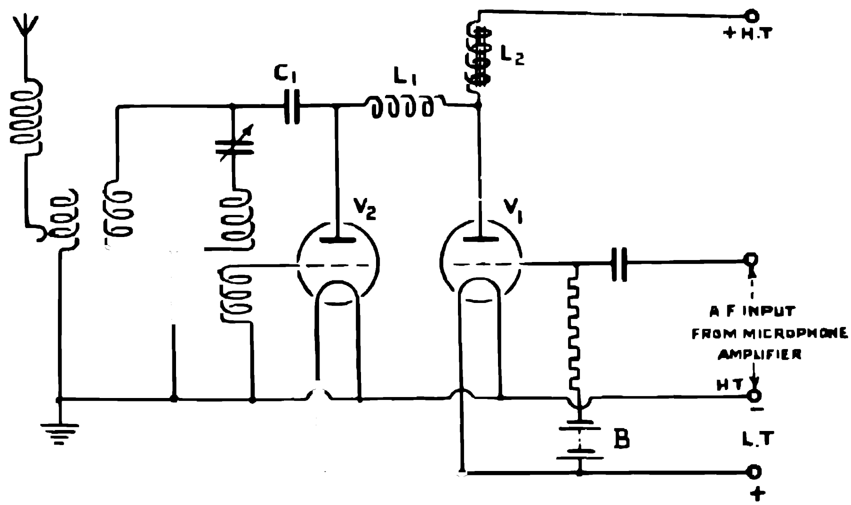
20. **Methods of Modulating.**—In the valve transmitter section (K.52), brief consideration was given to the fundamental principles underlying all methods of modulation, and it was observed that these principles apply equally to the production of type A3 waves. It is now proposed briefly to discuss certain special methods of particular importance in R/T work. These will include :—

(*a*) Two forms of anode modulation, namely, the HEISING OR CHOKE CONTROL method, and SERIES MODULATION.

(*b*) GRID MODULATION, both of the oscillator valve and of the amplifier.

(*c*) SUPPRESSOR GRID MODULATION.

21. " **Choke Control** " **Modulation.**—This method dates from about 1920 and is due to Heising, a typical circuit being shown in Fig. 17. V_2 is the R/F oscillator valve, coupled in the usual manner to an oscillatory circuit, which is tuned to the required carrier frequency. L_1 is a



CHOKE CONTROL METHOD OF MODULATING ANODE VOLTAGE

FIG. 17.

radio-frequency choke which serves to confine the R/F oscillations to the valve V_2 and its oscillatory circuit. V_1 is the modulator valve and L_2 is a large audio-frequency choke in the COMMON H.T. SUPPLY LEAD to both valves. A grid bias battery B serves to maintain the grid of the modulator valve working at the mid point of the straight portion of the valve characteristic on the negative side of the grid current characteristic, *i.e.*, class A conditions.

The audio-frequency modulating voltages from the microphone amplifier are applied between the grid and filament of the modulating valve V_1. Considered as part of the output circuit of V_1, the choke of L_2 is in parallel with the oscillator valve, and its impedance at audio-frequency is much greater (about 10 times) than the A.C. resistance r_a of the oscillator. This also applies to the blocking condenser C_1; the impedance of L_1 is negligible at audio-frequencies. The numerical value of the impedance of the output circuit of the modulator valve may thus be taken to be the A.C. resistance " r_a " of the oscillator valve. The modulator valve is essentially an audio-frequency power amplifier, and the audio-frequency voltage developed across its output, the oscillator valve, appears as a modulation of the anode voltage of the latter.

When the input voltage of the modulator valve is steady, the mean anode voltage of the oscillator, and hence the amplitude of the carrier, is also steady. The application of an audio-frequency input of the modulator valve produces a corresponding variation in the mean anode voltage of the oscillator, and therefore modulates the amplitude of the carrier in accordance with the original

microphone diaphragm vibrations. An increase of the anode current to valve V_1 produces a P.D. across L_2 (a back E.M.F.) which tends to reduce the anode current to both valves ; a decrease in anode current to valve V_1 would produce a back E.M.F. in the opposite direction. The A/F alterations in the current supply to valve V_2 are in anti-phase to the alterations in current to valve V_1, and it can be shown that the sum of the instantaneous currents is substantially constant. It is for the above reason that this method is sometimes known as the "constant current system of modulation."

Certain precautions are necessary in order to prevent distortion during this process. It was seen in Section "F" that, with the correct circuit conditions, the output resistance of a power amplifier stage should be twice the A.C. resistance of the valve to give the maximum undistorted output. Thus the A.C. resistance of the oscillator valve should be twice the A.C. resistance of the modulator valve. This involves the use of a valve of low A.C. resistance as modulator, or of two or more modulator valves in parallel. Alternatively, the choke may be replaced by a transformer with a suitable transformation ratio T. The effective resistance of the oscillator valve then becomes r_a/T^2, and so can be adjusted to the correct value.

The impedance of the choke L_2, or transformer, must be considerably larger than the A.C resistance of the oscillator valve at the lowest audio-frequency required. This choke must also be capable of carrying, without distortion, the direct component of the H.T. current, and superimposed on it a small A/F component of the modulator output current. By using a transformer the direct currents to the modulator and oscillator valves may be arranged to magnetise the core in opposite directions, and so to diminish its resultant steady magnetisation.

The percentage modulation which is obtainable in this way, depends upon the power output from the modulator. To produce 100 per cent. modulation, it is necessary to superimpose upon the D.C. anode voltage of the oscillator an alternating voltage big enough to swing the instantaneous anode voltage up to twice its steady value. Consider a numerical example :—

Let $L_2 = 2$ Henries, the H.T. supply voltage = 300 volts, and the mean anode current in $L_2 = 0{\cdot}08$ amps. If the maximum change in anode current is 20 per cent. at a modulating frequency of 1 kc./s., the maximum voltage change across $L_2 = (2\pi \times 1000) \times 2 \times 1/5 \times 0{\cdot}08 = 200$ volts. The effect of this is to vary the voltage across the oscillator valve between 100 and 500 volts.

If it is desired to increase the depth of modulation beyond the amount that can be obtained in the above circumstances, the modulator valve may be supplied with a higher H.T. voltage than the oscillator valve.

Attention has already been drawn to the increase in power output which occurs when the carrier wave radiated by a transmitter becomes modulated. It may be considered that the power required to produce the sideband components of a modulated wave is supplied to the oscillator from the output of the modulator.

This method of modulation is satisfactory if it may be considered that the output impedance is sensibly uniform over the range of audio-frequencies. This will be approximately the case provided that the impedance of L_2 is sufficiently large in comparison with the r_a of valve V_2. Departure from this condition introduces discrimination in the treatment of various frequencies, resulting in "frequency distortion." The chief limitation to the use of this type of modulation depends upon the fact that the frequency of an oscillator partly depends upon the anode voltage ; the effect of modulation is thus to cause the carrier frequency to vary, so introducing more frequency modulation. Choke control modulation is still used extensively in some of the older broadcast transmitters, but is now employed only where frequency stability is not essential. In some of the newer B.B.C. stations, it has been replaced by the modified choke control system, known as "series modulation."

22. Series Modulation.—Instead of the PARALLEL CONNECTED arrangement used in choke control modulation, in which an oscillator virtually constitutes the output impedance of the modulator, it is possible to arrange a SERIES CONNECTED circuit in which a driven H/F amplifier

serves as the load for the modulator; the stipulated conditions for choke control, therefore, still apply. Fig. 18 represents the essential circuit details of the system; a push-pull form of it is in use at the Droitwich B.B.C. station.

It will be seen that the modulator and amplifier valves are in series across the H.T. supply; the effect of variation in the A/F input to the modulator, working under Class A conditions, is to vary the series resistance of that part of the circuit, producing corresponding variations in H.T voltage between anode and filament of the modulated amplifier. It is now the usual practice to modulate the amplifier instead of the oscillator, in order to avoid the frequency distortion which the latter process involves.

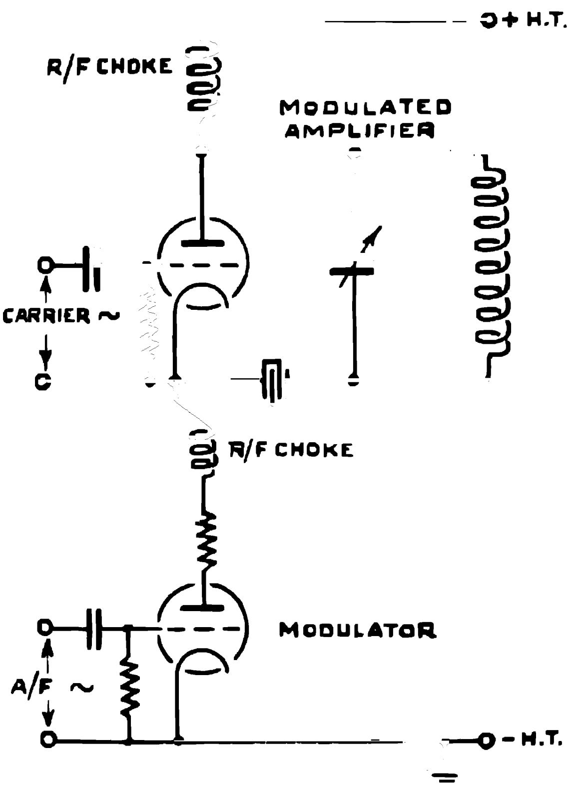


FIG. 18.

The filament of the modulator valve is at earth potential, while that of the amplifier valve and its associated input circuits are at a high positive potential above earth. This constitutes a difficulty inherent to this system, and it involves careful insulation of the source of filament heating current to that valve.

The system was introduced—

(a) after a decision to modulate at high power level, in order to avoid the difficulty of designing a long series of distortionless H/F amplifiers with band-pass characteristics (N.1); and

(b) to eliminate the use of the very expensive and bulky iron cored apparatus necessary when using "choke modulation" at high power levels.

It has been found to be a method of modulation singularly free from distortion, but it is important that the capacity to earth of the modulated amplifier circuits should be kept at a low value in order to avoid the shunting action which would occur at the higher modulating frequencies. The system suffers from the objection that a large value of H.T. voltage is required; it is usually about twice that employed in the choke control method. For this reason it may sometimes be more economically sound to employ the Heising method; a larger capital outlay is offset by an output stage of maximum efficiency.

23. **Grid Modulation of an Oscillator or Amplifier.**—A circuit of a grid modulated oscillator is shown in Fig. 19. The A/F output from the microphone amplifier is applied between grid and filament of the oscillator valve by means of a transformer coupling. The mean grid voltage is thus varied at audio-frequency, and the amplitude of the carrier is modulated accordingly. The condenser across the terminals of the transformer secondary acts as a by-pass for R/F and so ensures that practically all of the radio-frequency E.M.F. induced into coil B from the anode circuit is applied between grid and filament. In conjunction with the transformer secondary, it also provides a grid insulating condenser and leak resistance. The chief advantage of the system is the smaller

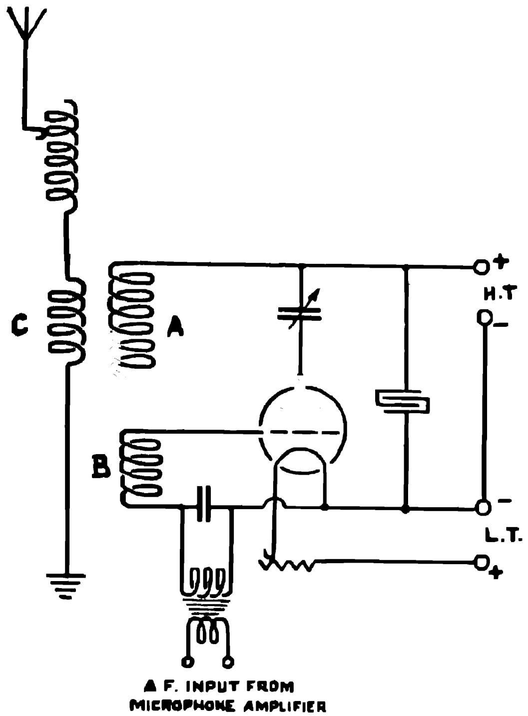


FIG. 19.

A/F power required for the same depth of modulation as in the choke control method, since the modulating voltage is applied to the grid of the oscillator. This economy in A/F apparatus may be somewhat offset by an increase in the number or size of the following R/F amplifying stages. A more serious objection is provided by the excessive amount of amplitude distortion which is introduced unless the percentage modulation is very low; for this reason, **simple grid modulation is now seldom used.**

In the case of a driven amplifier, R/T grid modulation may be effected in two ways :—

(*a*) Using a small radio-frequency carrier voltage and a large audio-frequency signal voltage, working under class A conditions with grid bias equal to half of the "cut-off" bias.

(*b*) Using a large radio-frequency carrier voltage and a smaller audio-frequency signal voltage, working under class C conditions, the amplitude of the carrier wave being somewhat less than the grid bias.

The method referred to in (*a*) is sometimes known as the van der Bijl type of modulated class A amplifier, and is the only one which is described here.

Fig. 20 (*a*) shows the circuit details by which an audio-frequency input may be superimposed upon the steady grid bias and applied between grid and filament in addition to the carrier frequency. The amplification of the small R/F carrier depends upon the position of the working point. As the working point alters with the A/F input, corresponding alterations will take place in the amplitude of the carrier, as shown in Fig. 20 (*b*). The percentage modulation will depend upon the curvature of the valve characteristic and the amplitude of the A/F input voltage; it will be 100 per cent. if the A/F input is large enough to operate the valve between the cut-off point and the point at which grid current begins to flow. This implies that the crest value of the A/F input should be approximately equal to the grid bias.

This type of grid modulation is very economical in power and easy to operate. It introduces a considerable amount of distortion, particularly at high depths of modulation.

It is important to realise the necessity for adjusting the working point to be somewhere on the curved part of the characteristic. If it were on the straight portion, the R/F carrier and A/F signal input would both be present in the grid circuit and continue to have separate existence in the anode circuit; the A/F input would not modulate the amplitude of the R/F carrier. Modulation consists in something more than mere ADDITION of two frequencies; it involves the MULTIPLICATION of frequencies. With reference to the work on sidebands (N.15), it will be recalled that the sum and difference terms, which represent the sidebands arising from the modulation, are due to "product" terms having the form B sin pt sin ωt.

The grid modulated class C amplifier produces modulation in a somewhat similar manner, and has the same advantages and disadvantages.

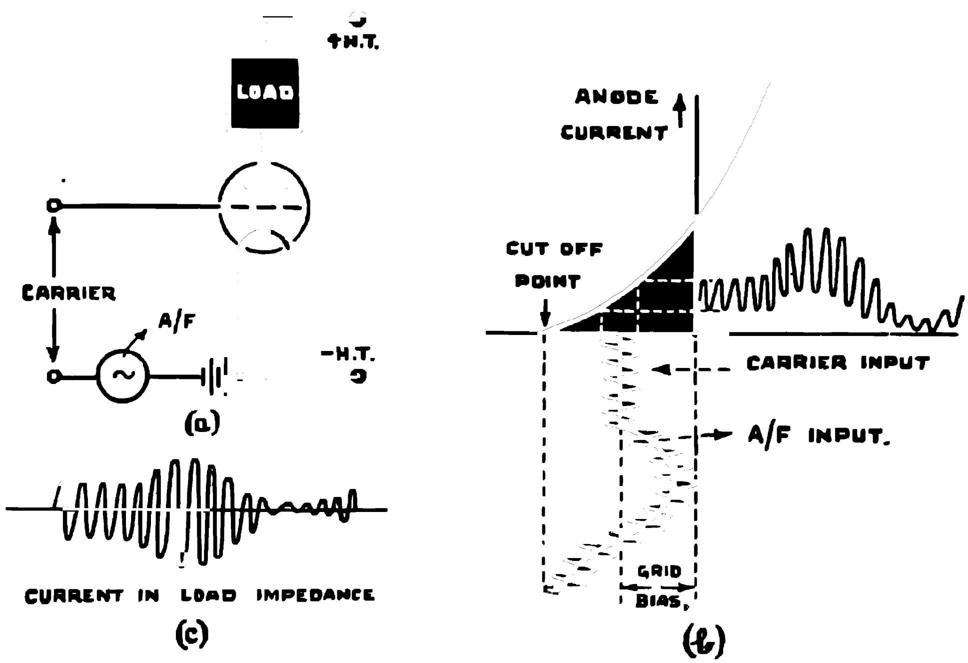


FIG. 20.

24. **Suppressor Grid Modulation.**—Where screen grid valves are used in H/F amplifying stages, it has been found possible to achieve modulation by varying the screen voltage. It is not a good method, since the distortion increases rapidly as the depth of modulation becomes greater.

Better results are obtainable using R/F pentodes. It is found that variations in potential of the "suppressor grid" produce corresponding alterations of output power, following almost a straight line law provided that the amplifying stage is suitably neutralised. This involves an alteration in the construction of the pentode. In normal use, the suppressor grid is connected internally to the cathode and is, therefore, earthed ; in this case that connection is broken, and the suppressor grid is made accessible by means of an external terminal.

Experiments in connection with suppressor grid modulation began in about 1933. It quickly became evident that the portion of its characteristic most useful for modulation purposes is in the negative grid bias region. Suppressor grid current flows when the grid is positive to the filament, and no current flows with negative suppressor voltage. Experiments also show that there is sometimes a reversal of suppressor current in the neighbourhood of zero voltage. For example, the current with the suppressor grid at (say) + 50 volts might be 1 milliamp ; at zero grid volts, the grid current in the opposite direction might amount to 20 microamps. This leads to the conclusion that over some portion of the characteristic the phenomenon of negative resistance is exhibited. In certain cases, therefore, care must be taken to have sufficient positive resistance in the circuit to prevent self-oscillation. As the suppressor grid is made more negative, corresponding changes take place in the amplitude of the R/F oscillations in the anode circuit. With an anode voltage of (say) 340 volts, a negative voltage of as much as 140 volts may sometimes be applied before the amplitude is reduced to zero, in other cases hundreds of volts may be required. For these reasons the suppressor grid is always biased negatively, though it swings positive on the modulation peaks. A small suppressor grid swing produces quite a large change of power in the output circuit.

The working point of the suppressor grid may be determined quite easily by experiment. It may be done by reading the anode current with zero grid bias (or the maximum safe positive bias) and then reducing the suppressor grid voltage, keeping the R/F carrier constant, until the anode

current is reduced to half its previous value. The grid bias in that state indicates the correct working point, and the A/F modulating voltages may be applied in series with the fixed suppressor grid bias, as shown in Fig. 21.

The R/F carrier input is applied between the control grid and filament, through the coupling condenser (1), supplied with the usual grid leak and R/F choke. The valve and the output impedance are fed in parallel with the usual R/F choke and anode blocking condenser. The filament connection to the centre of the output impedance, and condenser (3), enable the stage to be neutralised by the "tapped output" method. The screen grid bias is supplied from a tapping on the H.T. battery; it is fed through a small R/F choke, and condenser (2) maintains the screen and the filament at the same R/F potential. The suppressor grid negative voltage is supplied by a separate battery,

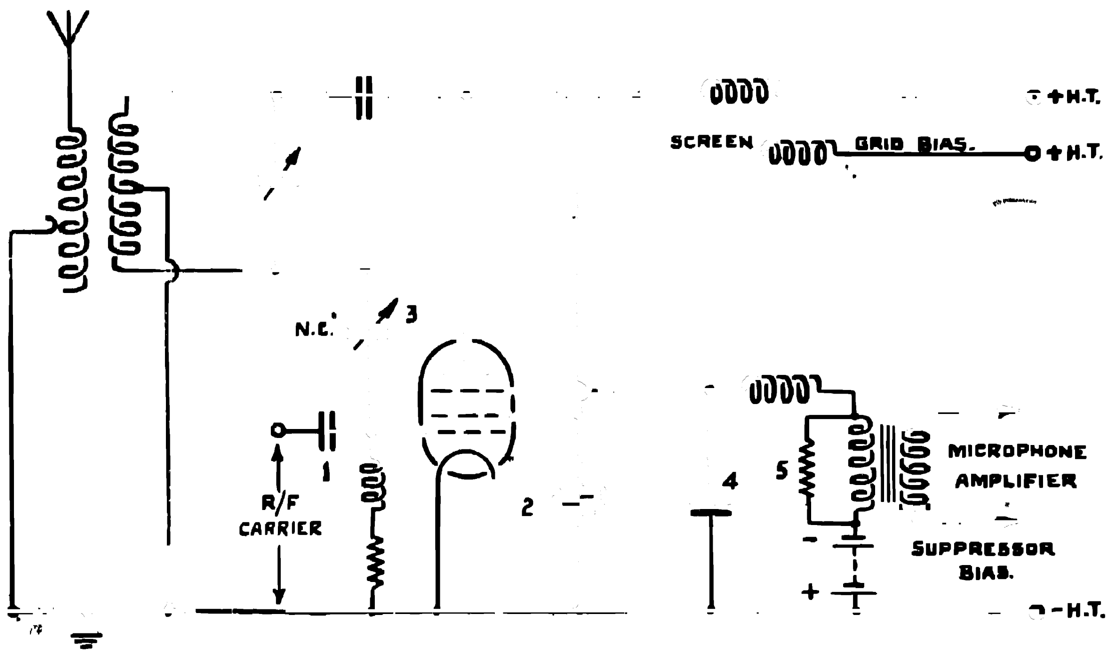


FIG. 21.

and is fed in series through the secondary of an A/F transformer and a small R/F choke. Condenser (4) serves to keep the suppressor grid and filament at the same R/F potential; in order to avoid attenuation of the higher modulating frequencies, care must be taken not to make this condenser too large. The resistance (5) tends to stabilise the load on the microphone amplifier. It is usually of the order of 4,000 ohms, and also serves to eliminate the possibility of self-oscillation arising from the production of negative resistance. The primary of this transformer is fed from the microphone amplifier.

Some practical values taken from a typical case show that with 350 volts positive on the anode and 150 volts positive bias on the screen, the optimum bias on the suppressor grid was about — 25 volts, the linear range about this point permitting an A/F input voltage of amplitude about 70 volts.

In comparison with "control grid" modulation, it will be noted that in this method the R/F carrier and the A/F input, are each applied between separate valve elements. In the older system, successful modulation depended upon the mutual adjustment of the two inputs and the working point; in the newer system, the requirements are almost independent. In certain cases, using neutralised stages, it is possible to obtain linear modulation up to practically 100 per cent., with almost negligible distortion.

It should be noted that the safe anode voltage using suppressor grid modulation is about twice that which is employed in the case of anode modulation. This is because the peak anode voltage is that of the D.C. supply in the case of suppressor modulation, but may amount to twice that value in the case of anode modulation. Considerable depths of modulation may be obtained using quite small modulating power.

25. **R/T Receivers.**—With the exception of the refinements already referred to in paragraph 1, the general procedure for R/T reception is the same as that for W/T, and receivers may be classified in a similar manner. There are three large groups which may be considered under the headings—(*a*) the straight set, (*b*) the superheterodyne receiver, and (*c*) the super-regenerative receiver.

Owing to the requirements of high amplification and selectivity, the number of examples of the first type of receiver has become a relatively small proportion of the whole. In a recent census of 22 commercial broadcast receivers produced by the same company, ranging from sets employing nine valves to those employing only three, it was found that only five of these fell into the category of "straight sets" (receivers with R/F amplification stages followed by a detector stage and A/F amplification stages), the remainder employing superheterodyne circuits. Examples of receivers employing the super-regenerative principle are rarely found, and are usually restricted to cases where high amplification combined with portability, are essential requirements; mobile police receivers are usually of this type.

In W/T reception, the main requirement is "**sensitivity,**" and implies a receiver capable of producing an audible noise of sufficient amplitude to enable the signal to be read. An R/T receiver, however, is not likely to be operated under conditions where sensitivity is a primary requisite, and the emphasis lies rather on the quality of the reception, *i.e.*, on the "**fidelity**" of the receiver; its audible output should be as faithful a reproduction as possible of the original sound vibrations which produce the modulation of the carrier. Both types of receiver will be required to possess a high degree of "**selectivity,**" the latter very much less than the former. Where the same receiver is to be used for all purposes, care must be taken that its fidelity is sufficiently good to give adequate reproduction of R/T. High quality R/T reproduction depends upon minimising the *distortion* of the modulated wave form. Since the object of any R/T receiver is to reproduce the A/F sounds corresponding to the modulation (or envelope) of the carrier wave, it is clear that care must be taken not to distort the envelope either in the R/F stages, the detector stage, or the subsequent A/F stages of the set. In practice, it is found to be comparatively simple to minimise distortion in aerial and R/F amplifier circuits, and the first source of appreciable distortion is likely to be the detector stage. It should be noted that receiver design usually begins here, since the requisite input voltage to this stage must largely determine the design of the preceding stages, the output similarly determining the design of the succeeding A/F stages. It is proposed briefly to consider the special requirements at the detector stage which characterise R/T receivers.

26. **The Perfect Detector—Linear Detection.**—The general principle of anode and cumulative grid detectors, as applied to W/T reception, are described in Section "D." In that account, no reference is made to the fidelity with which the rectified current represents the modulating frequency on the carrier wave. It is pointed out, however, that with the weak signals frequently obtained in W/T communication, using an anode bend detector, the rectified current is proportional to the *square* of the amplitude of the incoming signal voltage. For distortionless rectification, of course, an essential condition is that the rectified current should be proportional to the *first power* of the incoming signal voltage and not to its square. The rectified current produced by a **square law detector** is thus considerably distorted, and for satisfactory R/T detection a **linear detector** is necessary.

In Section "D" it has been observed that the perfect detector is a one-way device (a valve), possessing infinite resistance to applied voltages in one direction and no resistance to applied voltages in the opposite direction. If such a perfect rectifier were joined in a circuit with a source of E.M.F. but no external load resistance, with reference to Fig. 22, its "static characteristic" would consist

of the Y-axis above the origin, and the X-axis to (say) the left of the origin. In all practical cases, a load resistance is connected in series with the rectifier, which itself always has some resistance, in order that A/F variations in voltage may be developed across it and passed on to suitable amplifying stages. The " dynamic characteristic " of such an ideal rectifier, with a resistance in series, will be a straight line for all frequencies, if the resistance may be considered to be free from reactance.

Fig. 22 (*a*) represents the dynamic characteristic of a perfect detector. If the operating point is adjusted to the point O, where the sloping part of the characteristic meets the X-axis, the amplitude variation of the rectified current will be a replica of that of the input voltage for all values of the

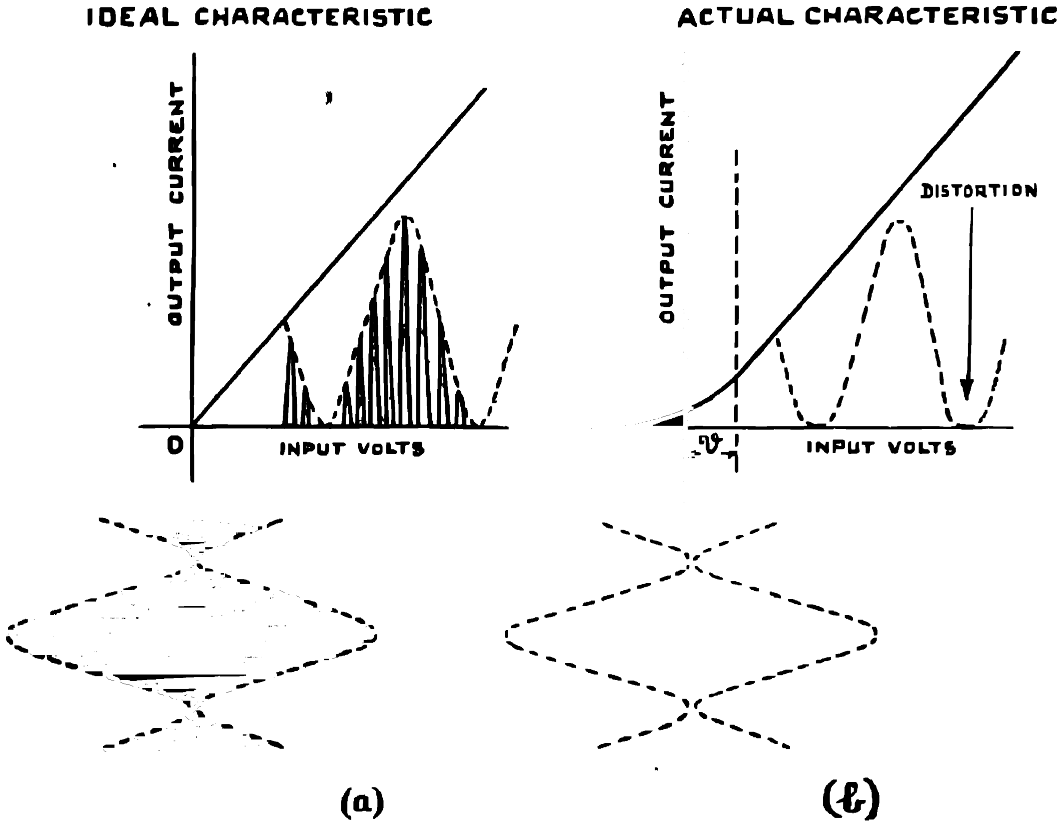


FIG. 22.

latter ; it will be observed that the figure has been drawn for 100 per cent. modulation. It may also be noted that where the output current passes through a pure resistive load, the shape of the wave form representing "output volts" will be the same as that representing output current.

In practice, there is no detector, depending for its action on thermionic emission, which provides a characteristic having a discontinuous slope of the kind shown in Fig. 22 (*a*). There are, however, various arrangements that make it possible to approximate in some measure to the characteristic of the perfect detector. For example, the dynamic characteristic of a triode can be arranged to have a fairly constant slope over a considerable portion of its length. The initial curvature shown in Fig. 22 (*b*) is, however, an essential feature of the characteristics of detectors depending upon electronic emission ; it represents the action of the space charge in limiting the anode current for small input voltages by repelling electrons back to the cathode. The curve implies that for small inputs the effective resistance is relatively high.

Somewhat similar characteristics are provided by the grid volts/grid current characteristic of a triode, and the anode characteristic of a diode, detector circuits based on the action of each being well known.

Detecting devices, having imperfect characteristics resembling Fig. 22 (*b*), are capable of giving linear detection, provided that the modulated input voltage amplitude never falls below the straight part of the characteristic. The feasibility of this depends, to a large extent, on the maximum depth of modulation of the carrier. With 100 per cent. modulation, the case represented in Fig. 22 (*b*), the amplitude varies between twice the amplitude of the unmodulated wave form and zero, and so it is impossible to avoid the curved part of the characteristic. The effect of the latter is to flatten the shape of the amplitude envelope of the output modulated wave form during the negative half cycles; this represents the introduction of distortion (N.7). From this it follows, that the greater the depth of modulation, the greater must be the mean amplitude of the input voltage, in order to ensure that the lowest input voltage amplitude reaches the straight part of the characteristic, having the value $\mathcal{V}$ of Fig. 22 (*b*). It is, of course, also necessary with larger input voltages to be certain that the maximum amplitude of the modulated input does not extend beyond the upper limit of the straight part of the characteristic, or into the region where grid current flows, in the case of anode bend rectification.

For the above reasons, detector stages in R/T receivers are always operated with relatively high inputs, and are, therefore, sometimes called "**power detectors.**" Moreover, since all modern R/T transmitters are capable of 100 per cent. modulation, every effort will be made to minimise the extent of the curved portion of the characteristic of the detector device in use.

It is proposed briefly to describe the essential features of three common methods of detection used particularly in R/T receivers, consisting of the two older methods sometimes known as "**power anode**" detection and "**power grid**" detection respectively, and the "**diode detector.**" The first two methods have recently been somewhat displaced in popularity by the diode detector; the latter is capable of giving almost distortionless detection and, in addition, is the basis of most methods of automatic gain control (A.G.C. or A.V.C.).

27. **The Diode Detector.**—The diode valve was one of the ealiest of the thermionic detectors, but even in the early days it received very little use; it was quickly succeeded by the triode which could detect and also amplify the signals after detection, and the use of a diode for detection was almost entirely discontinued. The simple diode is comparatively insensitive and suffers, moreover, from the disadvantage that its use involves the insertion of a comparatively low resistance in parallel with the tuned circuit preceding the detector stage; this introduces damping with consequent reduction in the selectivity of that stage.

The diode detector's return to popularity is largely due to the great increase in signal strength associated with modern broadcast transmitters. In many cases nowadays, R/F inputs to the detector stage amounting to 10 volts are available after one stage of R/F amplification. Further reasons for its return to favour lie in the insistent demand for distortionless detection using high modulation depths, and also for its use in producing the bias voltages of A.G.C. systems.

Fig. 23 (*a*) represents one form of the fundamental circuit of a diode detector, shown connected to a succeeding amplifying stage. The tuned LC circuit is joined in series with the diode and the load impedance R_1. Condenser C_1 functions as an R/F by-pass condenser shunting the load resistance. Common values of the load resistance vary from 0·5 megohms to about 2 megohms.

Fig. 23 (*b*) represents the anode characteristic of the valve. It should be noted that when the anode volts are zero, the anode current is not zero; with a load resistance in the circuit, the anode will hence take up a slightly negative potential, its numerical value depending upon the size of the load resistance. For example, with a load of 0·666 megohms a current of 1·5 microamps would produce a P.D. across the load resistance of 1·0 volts The *initial condition*, therefore, is represented by the point A, and OB is a "load line." If the load were to be increased to 1 megohm, the load line OB would be less steep, a current of only 1 microamp being required to produce an initial anode potential of — 1 volt.

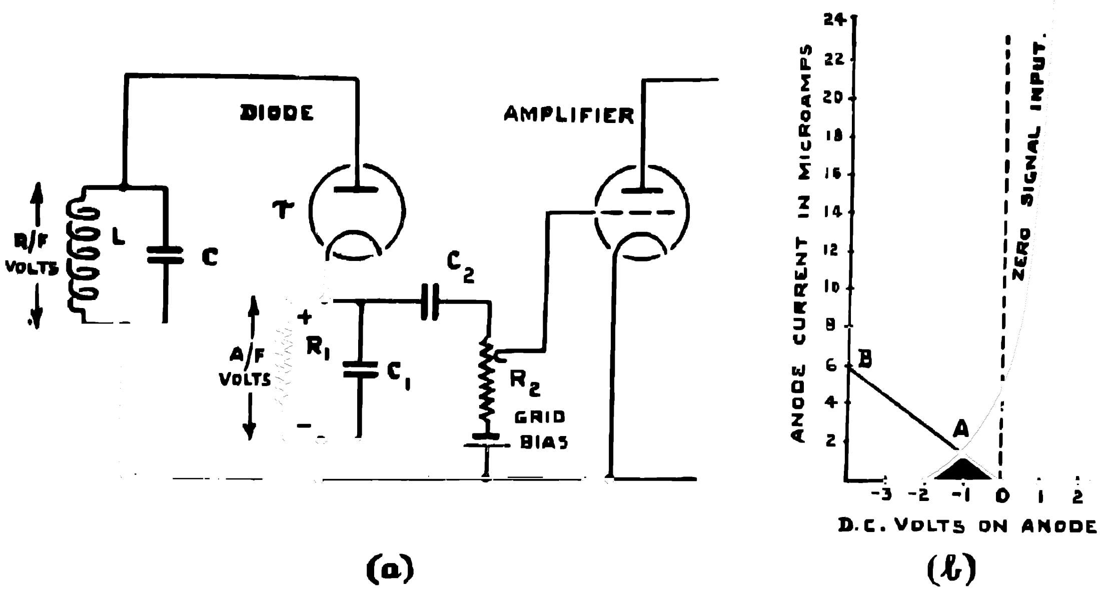


FIG. 23.

In order to appreciate the action of the circuit on receipt of an **unmodulated R/F input,** it is convenient for a moment to assume that the load resistance R_1 has been deleted from the circuit. We then have the conditions of a half wave rectifier on no load, or a peak voltmeter (H2) ; electrons flow during positive half cycles, and a P.D. will be developed across C_1 equal numerically to the peak value of the R/F input voltage. If the load resistance R_1 is then connected in shunt, a steady direct current will flow through it and a steady P.D will be produced across it. If R_1 is sufficiently big, the value of this P.D. will only be a little less than the peak value of the input voltage. For all inputs of value over a few volts, Fig. 23 (*b*) shows that this P.D. is strictly proportional to the amplitude of the signal input over a very wide range. An alternative explanation is to consider that the R/F input will produce a voltage swing about a working point given by (Fig. 23 (*b*)), the rise in current during the first positive half cycle being much greater than the fall during the succeeding negative half cycle. The mean current, therefore, rises and produces an increased negative potential on the anode as the load line OB is traversed. Succeeding complete waves repeat the process until the anode reaches an equilibrium potential, following which electrons only flow across the diode for a short time, at the peaks of the positive half cycles, in sufficient quantity to keep the condenser C_1 discharging at a constant rate. A diagrammatic representation of this process is given in Fig. 24 (*a*). It should be noted that detection of the R/F input has produced an output consisting of two components, a D.C. one, and an R/F one ; both have their uses.

In the case of a **modulated R/F input,** the D.C. potential difference developed across R_1 will rise and fall about its equilibrium value, and if the modulating frequency is an audible one, A/F variations of potential will be superimposed on the direct current P.D. By means of the circuit of Fig. 23 (*a*) and that of Fig. 25, these A/F variations of potential—the third component of the output—may be passed on to suitable amplifying stages. Alternatively, telephones may be inserted in place of the resistance R_1.

Fig. 24 (*b*) represents, diagrammatically, a method by which the characteristics of a diode may be investigated. Different values of the direct current P.D. developed across the resistance R_1 are provided by means of a battery and a low resistance potentiometer. The curves of Fig. 24 (*c*) are obtained by varying the above P.D., keeping the effective A/F signal voltage constant. The figure shows a family of these curves.

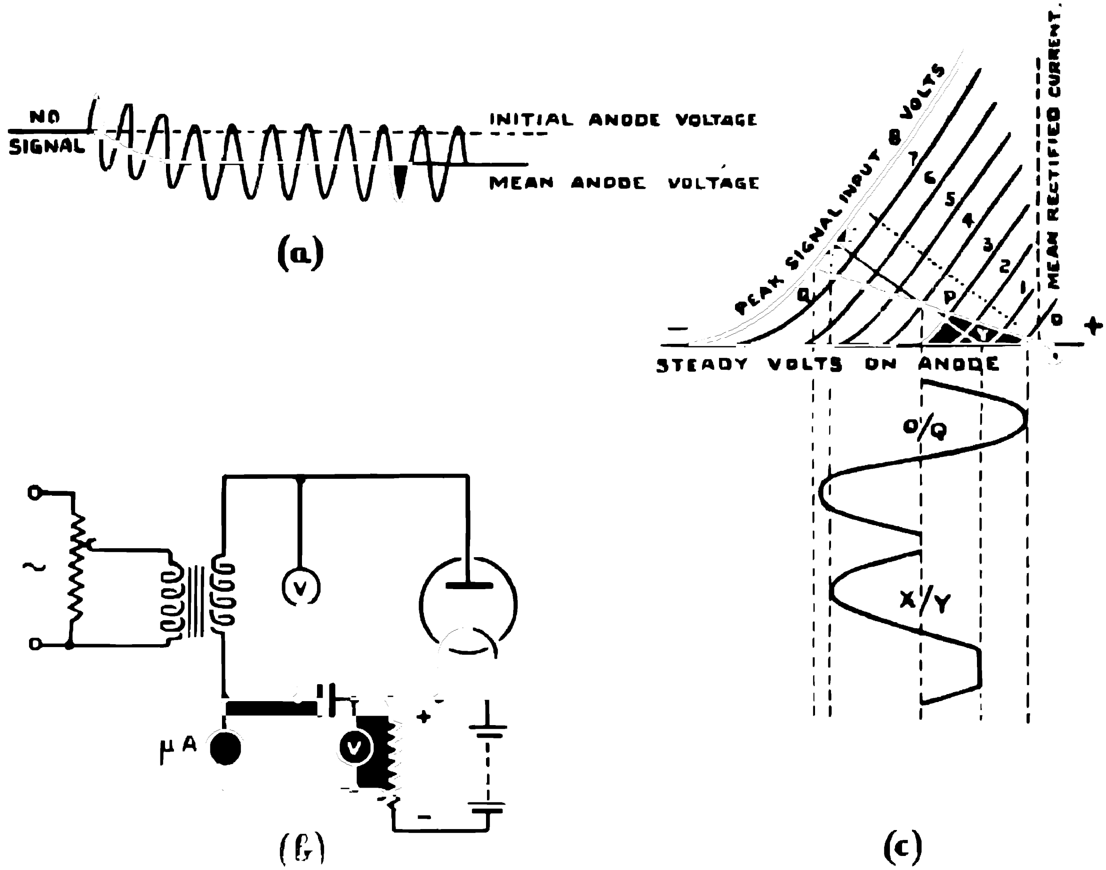


FIG. 24.

The line OQ represents a **D.C. load line.** For example, if R_1 is 0·5 megohms, a current of 10 microamps would correspond to a negative potential of – 5 volts. If the R/F input amounts to 4 volts, the working point is automatically set to the point P. If the input is a modulated signal, a voltage swing will take place about that point and, with 100 per cent. modulation, the load line will be traversed from O to Q. Under these conditions, the shape of the A/F wave form is almost a faithful replica of the amplitude envelope of the carrier signal—the curve O/Q ; the very small initial curvature of the zero signal input "line" introduces a little distortion on the modulation peaks.

In spite of the high degree of freedom from distortion, following almost linear detection, it is important to note that under certain conditions, the diode detector may give rise to serious distortion. When the diode is coupled to a succeeding A/F amplifying stage, it is usually necessary to prevent the D.C. bias potential from reaching the grid of the first valve of the amplifier. In Fig. 23 (*a*) this is achieved by means of condenser C_2, the reactance of which is sufficiently low at the audio frequencies concerned, but which offers an effective bar to the D.C. component. Resistance R_2 serves the dual purpose of grid leak and volume control. It will be noted that connection of C_2 and R_2 in parallel with the load resistance and C_1 must have the effect of reducing the overall impedance presented to the A/F component. In Fig. 24 (*c*) the load line XY represents the smaller load now presented to the A/F components. XY might be called the **A.C. load line,** since the circuit offers a different impedance to the two kinds of voltage. Unfortunately, this introduces a complication ; if an R/F signal, of magnitude 4 volts and having 100 per cent. modulation, is applied as before, the current in R_1 reaches zero when the R/F signal is in the neighbourhood of 1 volt. The effect of this is seen in the flattening of the waveform shown in the curve X/Y of Fig. 24 (*c*).

The problem of countering distortion due to this cause may be approached in at least two ways; the percentage modulation may be restricted to values not likely to cause distortion, or, alternatively, special networks may be used to make the impedance represented by the A.C. load line XY equal to that given by the D.C. load line OQ. In the next paragraph it is shown that the maximum modulation depth for no distortion is given by the relation $m = R_2/(R_1 + R_2)$. By making the size of R_2 three or four times the value R_1 it becomes possible to approach the figure of 100 per cent. modulation. Special networks are seldom used.

Another suggestion, due to Kirke, is to apply positive bias to the anode of the diode valve, in order to move the working point of the A.C. load line to the right, as shown by the dotted line in Fig. 24 (*c*); this is done so that the extension of this line should intersect the zero signal input line where it is cut by the D.C. load line. In practice, percentage modulations exceeding 90 per cent. are seldom used, and it is not possible completely to eliminate distortion, using 100 per cent. modulation, since one cannot completely straighten the initial bend which characterises all thermionic detectors. It can, however, be minimised by making the value of R_1 big in comparison with the effective resistance r of the rectifier. The A.C. resistance of the latter is equal to the reciprocal of the slope of the curve, dv/di, and the high value of that resistance at very low inputs is the cause of the bend at the bottom of the characteristic. It follows, therefore, that the smaller the value of the load resistance R_1 compared with the detector resistance at small inputs, the greater will be the extent of the bend. The relative value of these two resistances affects the question of " efficiency of rectification." The efficiency will be less at low inputs than at high ones, since an appreciable drop of volts will occur across the detector and reduce those available across the load resistance.

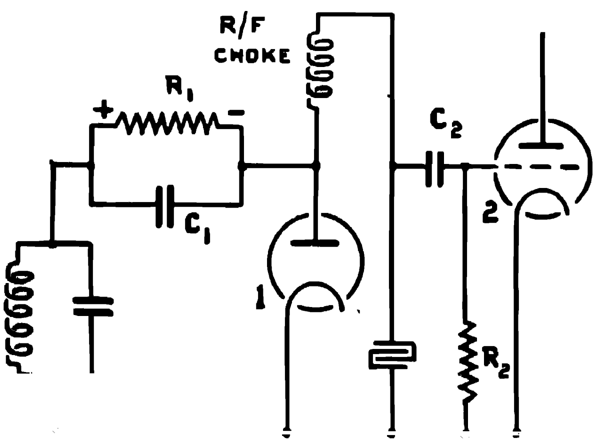


FIG. 25.

Fig. 25 represents another method of passing on the A/F variations in potential to an amplifying stage. Special care is taken to eliminate the R/F component by means of the R/F choke and by-pass condenser which are shown. It may be noted, that if direct connection were made to the grid of valve (2), the A/F variations in potential across R_1 would still be applied between grid and filament of valve (2), with the addition of the D.C. bias potential and the R/F component. Since both the grid of the triode and the anode of the diode consists of electrodes in the neighbourhood of a cathode emitting electrons, it is reasonable to suppose that there is no need to have both valves present. If the diode valve were removed, the circuit would become that of a grid current detector, the grid and filament functioning as the electrodes of the diode.

A method of straightening the initial bend in the characteristic of the diode (due to Kirke), consists in using a triode as a diode valve, the grid being maintained at a positive potential with regard to the filament, in order to neutralise some of the space charge which is responsible for the bend in the curve. Very satisfactory results have been obtained using polarising potentials of the order of +20 volts, and these circuits have become known as " Kirkifiers." Naturally, the grid current will be heavy, and one grave disadvantage of such circuits consists in the damping they introduce. The latter is sometimes equivalent to a parallel resistance as small as 8,000 ohms. For comparison, the parallel resistance associated with the ordinary diode may be shown to be roughly equal to $R_1/2$.

The value of condenser C_1 is chosen so that its reactance at radio-frequencies is small, and its reactance at audio-frequencies sufficiently constant over a given frequency range. To avoid "frequency distortion" it is essential that the impedance of the condenser and resistance combination should not fall at the higher modulating frequencies. It is shown below that this effect is minimised by using a value of C_1 of the order of $0 \cdot 0001\,\mu F$. Considerations already outlined have indicated the order of magnitude of R_1. Over-ruling all considerations, however, is the "CR value" or time constant of the combination; the latter must be adjusted to allow the superimposed A/F variations in potential to keep pace with the most rapid of those variations in the applied signal. A common CR value is $0 \cdot 0001$ seconds, but the older service W/T receivers use about twice this value.

In practice, the simple diode is seldom seen. It is usually found in combination with other valves within the same envelope, such as double diodes, duo-diode-triodes, etc. These valves feature in most modern broadcast receivers of the superheterodyne variety; extensive use is made of the D.C. component in the production of the negative bias of A.G.C. systems, to be described later.

★28. **Mathematical Note on the Diode Detector.**—(*a*) Fig. 26 shows the equivalent output circuit of the simple diode. The output circuit is required to present as constant an impedance as possible to the various *audible* modulating frequencies, for a given depth of modulation

The total impedance Z is given by—

$$\frac{1}{Z^2} = \frac{1}{R_1^2} + \frac{1}{X_1^2}, \qquad \text{and hence } Z = \frac{R_1 X_1}{\sqrt{R_1^2 + X_1^2}}$$

For a given A/F current of R.M.S. value I—

$$\text{A/F potential difference} = IZ = IR_1 \frac{X_1}{\sqrt{R_1^2 + X_1^2}}$$

The factor $X_1/\sqrt{R_1^2 + X_1^2}$ varies with the frequency and is always less than unity. For the best conditions, X_1 must always be much larger than R_1, and, in practice, X should not fall below about $2R_1$. If we take $0 \cdot 5$ megohms as the lowest normal value for R_1, an appropriate value for C_1 is found to be somewhat below $0 \cdot 0001\,\mu F$. Condensers of the latter value are, however, usually employed, resulting in a slight "high note" loss.

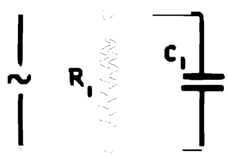


FIG. 26.

(*b*) The effective parallel load or "input resistance" due to a diode (or grid current detector) has been simply estimated by Terman.

Assuming that the valve only conducts on the *peaks* of the R/F signal input, the power absorbed is given by $\mathcal{V} i_{mean}$, where i_{mean} is the mean diode (or grid) current. Assuming the direct current P.D. developed across R_1 is *equal* to the peak signal input, and an "efficiency of rectification" β [where $\beta = R_1/(R_1 + r)$], we have (*cf.* H 4)—

$$i_{mean} = \frac{\mathcal{V}\beta}{R_1} \text{ ......and "Power absorbed" } = \frac{\mathcal{V}^2\beta}{R_1}$$

or in R.M.S. values .... Power absorbed $= V^2\left(\frac{2\beta}{R_1}\right)$

Hence the effective input resistance is $R = \frac{R_1}{2\beta}$

.... since "power" $= I^2R$ or $\frac{V^2}{R}$.

In many cases we may assume that $\beta = 1$.

(*c*) The case of the diode coupled to a succeeding A/F amplifying stage [Fig. 23 (*b*)] may be treated as follows:—

Assuming the value of C_1 to be chosen in the way outlined in (*a*) above, the effective load impedance presented to the A/F component is that represented by resistances R_1 and R_2 in parallel, since C_2 is an A/F by-pass condenser

Hence $\qquad Z \doteqdot R_1R_2/(R_1 + R_2) \qquad$ .... (1)

If the mean value of the direct current P.D. across R_1 is taken as $\mathcal{V}$, the superimposed A/F component due to the modulation may be written $N\mathcal{V} \sin pt$, where N is the percentage modulation (paragraph 14).

The peak value of the A/F current will be $N\mathcal{V}/Z$, and this cannot be greater than the mean direct current $\mathcal{V}/R_1$ .... (β assumed to be unity).

$$N\mathcal{V}/Z = \mathcal{V}/R_1$$

$$\therefore \quad N = \frac{Z}{R_1} = \frac{R_2}{R_1 + R_2} \ldots\ldots \text{from (1).}$$

This result gives the maximum value of N that can be handled without distortion. By making R_2 big in comparison with R_1 (3 or 4 times bigger) it is possible to approach full modulation, some distortion on the peaks being accepted. Clearly, other considerations may limit the size of the grid leak R_2 to not more than about 5 megohms.

29. **Power Anode Detection.**—In paragraph 26 it was observed that the dynamic characteristic of a triode resembles Fig. 22 (*b*). The use of the bends in that curve for the detection of weak signals has been fully described in Section "D"; in that case it is possible to assume a parabolic law for the curve, the use of which gives rise to the so-called "**square law**" detection, and a rectified output consisting of three principal terms, the D.C., R/F and A/F components.

In the case of power anode bend detection, the object is to obtain an approximation to "**linear detection**" for large R/F inputs. The available straight portion of the characteristic is limited by the lower bend, on the one hand, and the point at which grid current commences to flow, on the other. In order to make this linear range as large as possible, the detector stage is worked using a high anode voltage and a correspondingly large negative bias to adjust the working point to the middle of the lower bend. Under these conditions, the valve would be capable of giving almost distortionless detection of large input signals ranging between a minimum of value of $\mathcal{V}$ (*cf.* Fig. 22 (*b*)), and a maximum value corresponding to the point at which grid current would just begin to flow; the flow of appreciable grid current must be avoided in order to eliminate the consequent damping of the tuned H/F stage providing the input. In short, it is necessary to avoid both overloading or underloading the valve, in order to make the best use of the linear portion of the curve to the left of the point at which grid current begins, and thereby minimising distortion.

Owing to the difficulty of finding an equation to represent the action of the valve with large signal inputs, it is not possible to give any simple mathematical analysis comparable to that employed when the inputs are small and square law detection results. Although the full analysis is extremely complex, the practical results are very simple and are virtually the same as those resulting in the simpler case; the rectified output current contains the same three groups of components, the D.C., the R/F, and A/F ones.

If there is any appreciable distortion following a detector stage of this kind, it may be due to—

(*a*) Characteristic curvature of the so-called "straight line."

(*b*) Use of a percentage modulation approaching 100 per cent.

(*c*) The use of low inputs, such as might be expected from very distant R/T stations. In this case the distortion will be serious since the whole of the working range wil be non-linear.

In spite of these possible sources of distortion, anode bend detectors may be used with considerable success in broadcast R/T receivers.

The great advantage of this type of detector is the low damping it introduces into the previous tuned stage; when properly controlled, no grid current is taken and the equivalent parallel resistance may be as high as 200,000 ohms at broadcast frequencies. Moreover, when such a detector is fully loaded, its sensitivity is satisfactorily high.

As in the case of the diode detector, the extent of the initial curvature of the characteristic is minimised by using a relatively high output impedance. Preferably the latter should be a pure resistance, suitably shunted by an R/F by-pass condenser. The value of the latter is determined by the same considerations which obtain in the case of the diode detector; since it is usual to pick off the A/F variations in potential developed across the output resistance and to apply them between grid and filament of a succeeding A/F amplifying valve, it is convenient to arrange the by-pass condenser so that it may also act as a grid insulating condenser, preventing the application of H.T. to the grid of the A/F amplifier. Employing the usual terminology, the detector valve is resistance-capacity coupled to the succeeding A/F stage.

Finally, the question of R/F feed-back from the output circuit to the input through the anode-grid inter-electrode capacity must be considered. In this case, the R/F output circuit has a nett capacitive reactance, and so the feed-back is in such a phase as to damp the input circuit (Section " D "). This effect is the same as " reaction " but opposite in phase, and is sometimes called " **anti-reaction** " ; analysis of this feed-back of energy from anode to grid of the valve was first made by J. M. Miller, so giving rise to the name " **Miller effect.**" The feed-back increases with the reactance of the output condenser, *i.e.*, it varies inversely as the capacity, and from this point of view the capacity should be the largest possible.

It is fairly obvious that general rules cannot be laid down for the best values of C and R under all conditions. They must be chosen with respect to the individual circuit, the type of valve employed, and so on. As an instance, the figures in one broadcast receiver, whose detector valve constants were $r_a = 25{,}000$ ohms, $m = 25$, were R = 250,000 ohms and C = 0·0003 μF. (*cf.* paragraph 27, for CR values).

30. **Power Grid Detection.**—In Section " D," and in N.27, it was shown that a grid current detector may be regarded as a diode detector followed by a triode amplifier of the resulting A/F variations in potential developed across the load resistance. In this case the latter function is performed by the grid leak, and the rectified current is now grid current. The grid condenser and leak resistance thus play similar parts to the by-pass condenser and load resistance in the case of the diode detector and anode bend detector circuits respectively. Their values will be dictated by considerations similar to those already discussed. The I_g/V_g characteristic is like that of a diode, and has less curvature than the static mutual characteristic ; the extent of the curvature may, therefore, be limited by using a ratio of leak resistance to grid/filament working A.C. resistance which is not so high as for the corresponding case in anode detection. This allows the use of a larger relative value of capacity without exceeding a stipulated CR value.

On application of an unmodulated R/F signal, the grid runs negative to a value approximately equal to the peak value of the signal voltage input ; the working point is moved to the left and the anode current falls. If the carrier is modulated, the position of the working point will vary in a manner which produces anode current waveform variations which are a faithful replica of the modulating wave frequencies. In view of the larger grid input voltages, the H.T voltage on the anode must be sufficiently high to prevent the greatest signal amplitudes during the negative half cycles from reaching the lower bend of the mutual characteristic in use. If this is not done, anode bend detection will take place, tending to *raise* the anode current, at the same instant of time that grid current detection is occurring, tending to *reduce* the anode current ; the nett result will be to produce distortion, an effect which may be countered by using high anode voltages, or by performing the functions of detection and amplification in separate valves, as in the case of the circuit of Fig. 25. Overloading of a grid detector is a commonplace occurrence.

The detector stage is usually followed by a stage of A/F amplification, to which it may be coupled by any of the usual methods. Resistance-capacity coupling gives the best results, but requires a high H.T. voltage in order to give a high enough anode voltage after the drop in the resistance has been taken into account.

In order to emphasise the difference in the operating conditions between power grid detection and ordinary grid current detection, as used in W/T practice, the following comparative table of values is given :—

——	Anode Voltage.	Condenser.	Resistance.
		μF.	Megohms
Cumulative grid	50	0·0001	2·0
Power grid	150	0·0003	0·2

The most serious disadvantage of power grid detection is the damping it produces on the input circuit, which is much greater than in the case of the power anode detector. It arises through two causes :—

(a) The flow of grid current necessary for detection, involves a diminution of the working A.C. grid/filament resistance. This is in parallel with the input circuit, and so its damping effect is inversely proportional to its actual value.

(b) The damping due to Miller effect, i.e. the feeding-back of energy through the anode/grid capacity in anti-phase to that fed back by "reaction" circuits, is greater than in the case of the anode bend detector. This is due to the fact that in this case the operating part of the mutual characteristic is the straight part, where the amplification factor for the R/F input has its highest value. It is much higher than the effective amplification factor for a valve working on the lower bend. The amount of feed-back obviously depends on the voltage developed across the output circuit, and so is proportional to the amplification factor for given values of inter-electrode capacity and R/F output impedance.

The higher amplification given by power grid detection is more than sufficient to compensate for this increased damping of the input as compared with power anode detection, but a decrease in selectivity is inevitable.

The damping due to feed-back (Miller effect), may be considerably reduced in both power anode and grid detection, by using two valves in the detector stage, their grids being in push-pull and their anodes in parallel. Alternatively, it may be countered by the use of neutralising circuits. The use of a screen grid valve for detection would also minimise feed-back in both cases, as has already been seen when discussing amplifier stability (Section " F "). The ordinary screen grid valve characteristic is, however, unsuitable, though a push-pull screen grid valve detection stage has been successfully used.

In general, power grid detection leads to less distortion in the rectified output than power anode detection. For some considerable time, power grid detection was the standard method in the highest quality receivers; to some extent it is now falling into disuse, owing to the higher field strengths which are now usually available, and the consequent suitability of a diode for the same purpose.

31. **Aerial and R/F Amplifier Circuits.**—In W/T reception, provided audibility is ensured, the chief consideration in the design of the aerial circuit, and the R/F amplification stages, is the provision of sufficient selectivity to enable the desired signal to be received with the minimum of interference. It has been seen that this may be accomplished by the use of an aerial secondary circuit—often untuned—with loose coupling to sharply tuned circuits in the subsequent R/F stages. In R/F reception, selectivity is also desirable, but its attainment conflicts with another important quality, namely, that the response of the receiver should be uniform over the band of frequencies representing the sidebands and the carrier. If the circuit is too selective this will not be the case, and the higher audio frequencies in the final output will be relatively weaker than the lower ones, thus giving rise to the form of distortion said to be due to "sideband cutting."

For undistorted R/T reception, the response of the receiver should be uniform over a band of frequencies of width equal to twice the highest modulation frequency. This is very different from the response of a sharply tuned circuit, in which maximum sensitivity is obtained at one selected frequency, the response falling off rapidly on departure from it—Fig. 27 (a). It is obvious that, if the circuit is tuned to the carrier, the distortion is greater the more the sideband frequency differs from the carrier frequency. In other words, attenuation of the higher audio frequencies is the price of high selectivity. A compromise has therefore to be made.

32. **Band-Pass Filters.**—The ideal R/F response curve for an R/T receiver is shown in Fig. 27 (a), for comparison with the response curves of a sharply and flatly tuned circuit respectively. It is possible to approximate to this ideal curve by making use of the double frequency effect produced in two circuits tuned to the same frequency and coupled to each other (Volume I). It has been seen

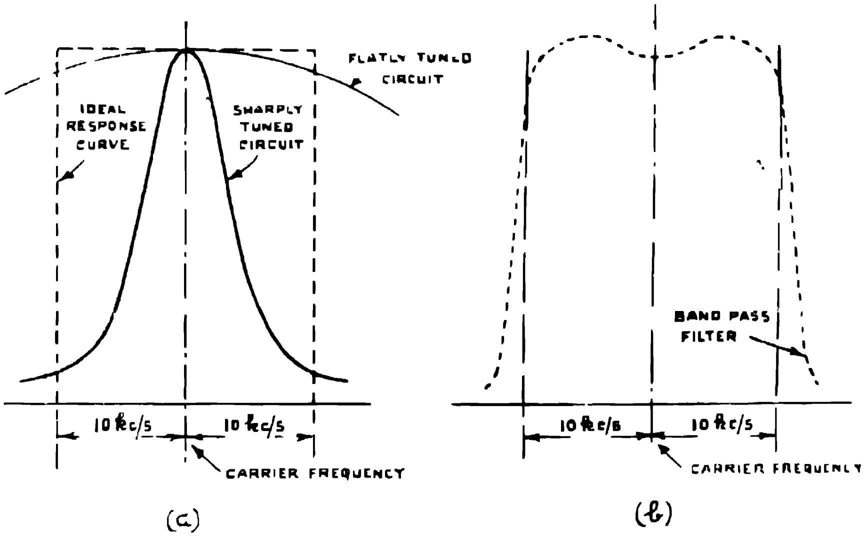


FIG. 27.

that two peaks in the response curve are then produced, their distance apart increasing as the coupling is made tighter.

The primary and secondary circuits are separately tuned to the carrier frequency and the coupling is arranged so that the resonance peaks are each about 5 kc /s. from the carrier frequency, giving a response curve as shown in Fig. 27 (*b*). When used in this way, the coupled circuits form a **Band-Pass Filter,** since they respond nearly uniformly to a band of frequencies, and hardly at all to frequencies outside the band. Provided the interfering frequency is more than about 10 kc./s. removed from the carrier frequency, sufficient selectivity as well as a fairly uniform response to the modulated carrier is obtained. With a constant coupling factor, the separation of the peaks is always the same percentage of the carrier frequency, and so increases with that frequency. But the band-pass filter is required to give a constant peak separation over a range of carrier frequencies. This may be obtained by a filter employing a " mixed coupling," or by means of a circuit incorporating " variable selectivity " as one of its features ; it enables the band-pass width to be adjusted to suit the frequency.

An alternative method is to cut out interference by using very sharply tuned circuits in the R/F stages, compensating for the resulting sideband attenuation by means of a **tone corrector** in the A/F stages. The A/F output circuits are designed so that they give a greater amplification of the higher audio-frequencies, in order to minimise the effect of sideband cutting, which involves a smaller response in the R/F stages to the sidebands corresponding to the higher modulating frequencies. The rising characteristic of the A/F tone corrector circuit is followed by a sharp cut-off at about 8 kc./s , and so gives approximately the same result as a band-pass filter in the R/F stages.

A form of interference which may sometimes be observed is known as " **sideband splash** "; it is due to heterodyne action between the carrier of the *wanted* station and interfering sideband frequencies of an *unwanted* station. The effect can only be countered by an increase of selectivity.

33. **Volume Control. Back Coupling. Cross Modulation.**—The number of R/F stages required depends on the type of detection employed, a power anode detector requiring a higher input voltage than a power grid detector, and a diode, used also for the purposes of A.G.C., requiring more than either of them. It is important to maintain this input voltage at its best value for distortionless detection, and sufficient R/F amplification must be available for this purpose with the most distant signals it is desired to receive, some form of volume control being incorporated in the first stage to prevent overloading with stronger signals. Volume control systems are now usually automatic in action, although a manual control is always included.

To obtain high amplification, screen grid valves or H/F pentodes—usually of the variable mu type—are commonly used in the R/F amplification stages ; their employment avoids the neutralisation necessary with triodes. The usual precautions taken against back coupling in W/T amplification stages, are even more essential in R/T receivers ; even if the stray coupling is not sufficient to give rise to self-oscillation, it may produce distortion.

Ordinary screen grid valve dynamic characteristics are not so straight as those of triodes and distortion may arise, either due to partial rectification of the wanted signal, or to rectification of an interfering signal which then modulates the carrier of the wanted one. This latter effect is known as **cross modulation,** and may be minimised by the use of a band-pass filter in the aerial circuit and by the use of variable mu valves.

The phenomenon is usually observed when there is a neighbouring strong *unwanted* station, capable of applying an input voltage sufficiently great to cause appreciable rectification if an ordinary triode or screen grid valve is in use in the first R/F stage. The rectified current will then appear as a modulation on the carrier of the wanted station, and both stations will be heard ; the unwanted station usually disappears if the wanted station ceases to radiate its carrier, since the tuned circuits subsequent to the first R/F stage are sufficiently selective to reject the unwanted frequency. The possibility of cross modulation would be increased, if, for any reason, the working point of the first R/F stage were to be set nearer the cut-off point ; an early form of volume control depended on the reduction of g_m obtained by increasing the negative grid bias of an ordinary triode or SG valve in the first R/F stage.

Variable mu tetrodes and pentodes have mutual characteristics possessing a long "tail" of small and slowly decreasing slope, instead of the usual sharply curved lower bend. The slowly decreasing value of mutual conductance (g_m) makes them admirable for use in A.G.C. circuits, which depend for their operation on a smooth control of the stage gain of the initial R/F amplification stage or stages. An enormously wide range of volume control is obtainable by means of a variable grid bias adjustment. Moreover, the small curvature of the characteristic associated with low values of g_m, renders these valves highly suitable to deal with unwanted large input signals, without the consequent detection followed by cross modulation which would be produced by an ordinary valve.

Typical characteristics of variable mu screen grid valves and R/F pentodes are shown in Fig. 28 (*a*).

34. **Manual Volume Control.**—There are several simple ways of hand controlling the R/F input to the detector stage One method involves the use of a high resistance potentiometer across the aerial coupling coil, or aerial condenser. The aerial is then taken to a sliding tap on the potentiometer, the latter being varied as requisite in order to control the input from the aerial. It will be remembered that the damping effect of a shunting resistance is inversely proportional to its actual value, and if the potentiometer is of high enough resistance its damping effect is negligible. With this method of control, the R/F amplifier always works at full gain, a state of affairs which is undesirable while listening to a strong R/T station, on account of the background noise (valve hiss) which is usually present.

Alternatively, the magnification of the tuned circuit providing the input to the first R/F amplifying stage may be directly controlled by means of a variable resistance joined in parallel with the former. When the parallel resistance is small, the tuned circuit is highly damped and the gain is much reduced. The method is little used, its great disadvantage being the reduction of selectivity which it produces.

The use of variable reaction is a very old method of providing manual control of the volume. The disadvantage of reaction is that it sharpens the tuning, the selectivity increases, and its use is liable to give rise to distortion produced by "sideband cutting."

A fourth method consists in controlling the stage gain of a screen grid R/F amplifying stage by varying the positive bias potential of the screen. Inspection of the mutual characteristics of a screen grid valve shows that, for low values of the screen bias, the slope of the appropriate curve is low, and the voltage amplification factor of the stage correspondingly small. For higher values of screen bias, the V.A.F. is greater. The method is open to the objection that the curvature of the characteristic and the sharpness of the cut-off make it impossible for the valve to handle anything other than signals of very small magnitude.

The most satisfactory method of volume control is provided by the use of variable mu screen grid or R/F pentode valves. Typical characteristics are shown in Fig. 28 (*a*); the substitution of a gradual change of slope for the usual sharp lower bend, indicates a delicate method of controlling the gain of an R/F amplifying stage. The method consists in varying the working point of the valve by means of an adjustable negative bias. It is sometimes claimed that the variation of slope of the characteristic follows a logarithmic law, the effective stage gain varying correspondingly. Since the human ear responds logarithmically to sounds of different intensities (Appendix A), the audible variation of the volume is smooth and uniform. Signals up to about 3 volts (peak) may be handled without serious distortion due to detection caused by characteristic curvature.

35. **Automatic Volume Control.**—With the growth of R/T beam transmission using H/F, and with the increasing interest taken in the reception of M/F broadcast transmissions at distances beyond the limits of the "service area," some method of overcoming the disadvantages of fading became necessary. Manual volume controls became too troublesome, and an automatic device was needed which would maintain the volume approximately constant at any desired level. With headphones, or with a small receiver in an ordinary room, a fair amount of fading can be tolerated; but whenever the programme is reproduced at power through large loudspeaker systems, quite a small amount of fading is sufficient to spoil the programme value.

The possibility of automatic volume control—hereafter usually called automatic gain control (or A.G.C.)—depends upon the use of linear detectors capable of dealing with input signals of large magnitude, such as diodes or metal rectifiers, in which case it has been shown (paragraph 27) that the magnitude of the D.C. component is strictly proportional to the peak value of the carrier input; it may be said that the principle of A.G.C. lies in the use of the D.C. component (Fig. 23) of the rectified current to produce the necessary negative bias, required to control the *gain* of the preceding variable mu R/F amplification stage or stages, and sometimes that of the frequency changer stage. Under suitable conditions, it is possible so to control the gain that wide variations of incoming signal strength result in only small variations of the detector input.

Care must be taken not to assume that A.G.C. is a complete cure for fading; it is only a palliative, and its use is attended by certain drawbacks. For example, when bad fading is taking place the background noise will be increased in intensity until it is as loud as the original signal. Moreover, under conditions during which the carrier and sideband frequencies do not fade out simultaneously (selective fading P.14), during a period when the carrier may be very weak the sidebands may be relatively much stronger, and would be passed through the receiver (owing to the dependence of A.G.C. on the carrier) at a greater strength than when the carrier frequency is present. The result is a periodic blast of harsh, high-pitched reproduction, and it may frequently be preferable to limit the action of the A.G.C. device so as to let fading occur after the signal strength drops below a certain value.

In commercial practice, there are numerous forms of automatic volume control, among which the following are well known:—

SIMPLE A.G.C.—In this case, with zero signal input, all amplification stages have maximum sensitivity, but an A.G.C. bias is produced and fed backwards with all signals, even the smallest.

DELAYED A.G.C.—This involves the use of a special circuit in which the A.G.C. bias does not appear until the R/F input to the detector reaches a certain pre-arranged value; this results in the sensitivity of the receiver remaining at a maximum for weak signals.

AMPLIFIED DELAYED A.G.C.—This is used when it is desired to minimise the R/F amplifying stages receding the detector. The necessary large A.G.C. bias is obtained by use of the principle of a "direct current amplifier."

CORRECTED A.G.C. WITH DELAY.—With the ordinary forms of A.G.C., a very large change in signal strength reaching the aerial is *almost* completely compensated for by the resulting change in amplification. Some change in input to the detector is, however, necessary in order to produce the compensating change in amplification.

It is possible to correct almost completely for the latter by applying a portion of the A.G.C. bias to a variable mu pentode in the A/F amplifier; nearly perfect A.G.C. can be produced in this way, no change in A/F output resulting from any change—within wide limits—in the strength of the input to the receiver.

QUIESCENT AUTOMATIC VOLUME CONTROL (Q.A.V.C. OR Q.A.G.C.).—This is a modification of the simpler forms, and its object is to produce quiet inter-station tuning, eliminating the background hiss and "set noise" which is heard when the receiver is working in its state of maximum sensitivity.

VOICE OPERATED A.G.C. CIRCUITS.—These are of importance in public address systems, where it is desired that the volume of reproduction should be independent of changes in the relative loudness of the announcer's voice.

It is proposed briefly to consider certain of these forms of A.G.C.

36. **Simple A.G.C.**—Fig. 28 (*b*) represents the circuit details of a variable mu R/F or I/F pentode amplifying stage (1) followed by a diode detector (2). It will be seen that the diode detector part of the circuit is similar to that of Fig. 23, R_1 being the load resistance across which is developed the direct current P.D. capable of supplying a negative bias equal approximately to the

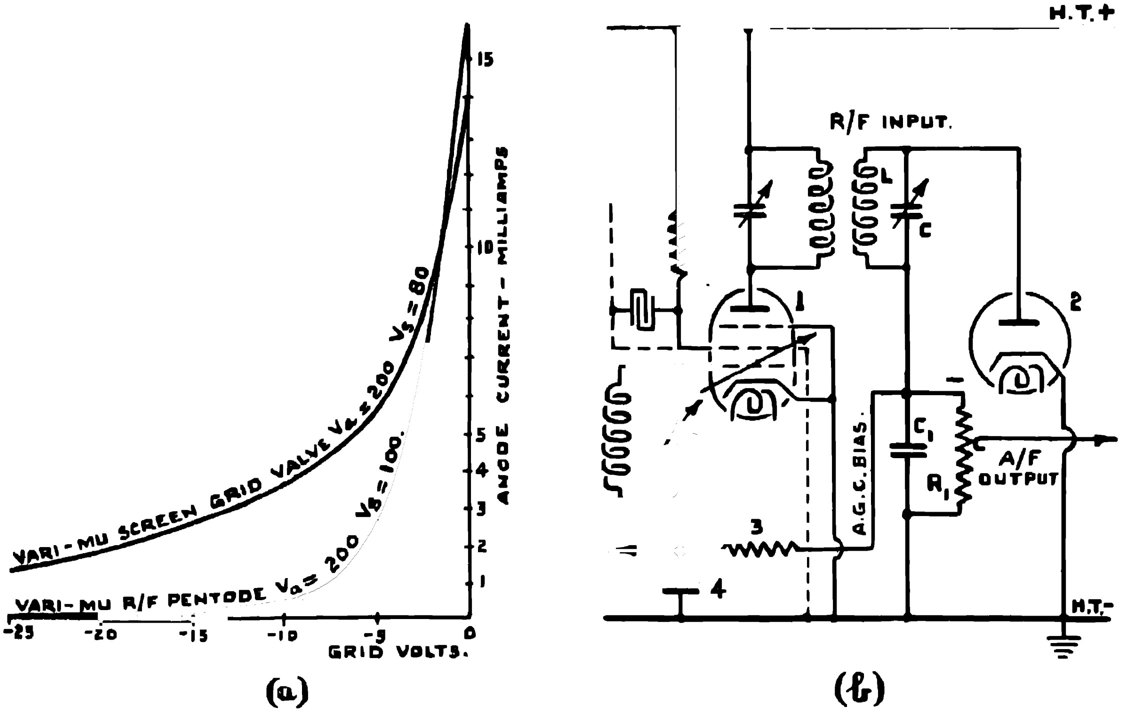


FIG. 28.

peak value of the input signal volts. The negative bias is fed back through a low pass filter to the grid of the variable mu pentode. A low pass filter is necessary to pass only the D.C. bias due to the relatively slow changes in the amplitude of the carrier wave, and to prevent the A/F changes due to modulation being fed back to the amplifying valves. Thus, a small increase in the strength of the carrier frequency reaching the detector results in an increase in the rectified voltage; this produces increased negative grid bias for the amplifying valves, and reduces the amplification to an extent which *partially* compensates for the increase in signal input which gave rise to the above change.

The low pass filter consists of the resistance (3) and condenser (4) shown in the diagram. In most cases the value of the resistance is about 1 megohm, and the size of the condenser is adjusted so that the filter will pass to the grids of the amplifying valves all frequencies below about 10 cycles per second, but will obstruct all frequencies above that value. A suitable size for this capacity is 0·05 microfarads. In this way, only the slow changes in D.C. potential reach the grids of the amplifying valves from the detector stage, the more rapid A/F changes being obstructed by the filter. Owing to the high value of the filter resistance, it is essential that the whole A.G.C. circuit should be particularly well insulated from earth, only the best components being used. The resistance to earth should always exceed 20 megohms.

Fig. 28 (*a*) represents the mutual characteristics of a Service type of variable mu screen grid valve and R/F pentode respectively. The smoothness of the curves gives an indication of the steady reduction of amplification which is obtainable as the A.G.C negative bias is steadily increased.

The A/F variations in P.D. due to the modulation are passed on to subsequent A/F amplifying stages through a manual volume control, is indicated.

The immediate reduction of sensitivity of the receiver on receipt of even the smallest carrier wave, constitutes one of the more serious objections to this simple form of A.G.C. The system is, therefore, not good for the reception of weak signals Other objections include :—

(*a*) The presence of a noisy background when signals are fading badly.

(*b*) The noisy inter-station tuning—hence the need for Q.A.G.C.

(*c*) The slightly increased tuning difficulty, owing to the tendency for the A/F output to remain constant even when slightly detuned. This fact points out the advantage of "tuning indicators" and "automatic tuning" devices, now commonly fitted to the more expensive receivers. In its simplest form, the tuning indicator could consist of a milliammeter in the anode circuit of one of the controlled stages; the circuit will be in tune when the carrier frequency wave has its maximum amplitude, producing in turn the largest A.G.C. negative bias and the smallest value of the mean anode current.

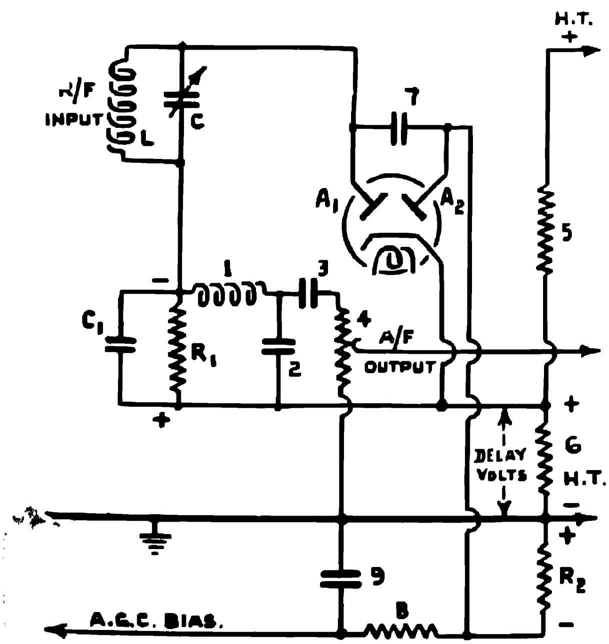


FIG. 29.

37. **Delayed A.G.C.**—The undesirable decrease of sensitivity on receipt of even the weakest signal—a salient feature of simple A.G.C.—may be overcome by arranging to delay the production of A.G.C. negative bias until the input to the detector reaches a certain pre-arranged value. This may be done by means of the circuit of Fig. 29. The valve is a double or (duo) diode, and in this case the two diodes are used in parallel, one being used to detect the signal frequency and the other to produce the A.G.C. negative bias. The signal frequency is received by the tuned LC circuit and fed between Anode A_1 and cathode, through the R/F by-pass condenser C_1, having the load-resistance R_1 in parallel with it. The D.C. component develops the usual negative potential at one end of resistance R_1, and the A/F variations in potential are fed through an R/F filter circuit consisting of inductance (1) and condenser (2), and applied between grid and cathode of a succeeding A/F amplifying stage.

Condenser (3) is an A/F by-pass condenser, and serves to prevent the application of negative bias to the A/F stage; resistance (4) constitutes the manual volume control.

Resistances (5) and (6) form a potentiometer across the H.T. supply, the negative terminal being connected to the earth line. This has the effect of making the cathode positive with respect to anode A_2, which is earthed through resistance R_2; the value of the positive potential above earth is equal to the drop in volts across resistance (6).

The signal input is also applied between anode A_2 and cathode, by means of condenser (7). Since, however, the cathode is positive with respect to A_2, no current will flow in R_2 until the peak value of the signal input exceeds the value of the "delay volts" available across resistance (6). The delay can, therefore, be precisely determined, and when it has been overcome, current will flow in R_2, and the resulting negative bias may be fed backwards to the R/F amplifying stages. Resistance (8) and condenser (9) constitute the usual low pass filter.

The method has the disadvantage that considerable R/F amplification must precede the stage. It is, however, better than the simple A.G.C. system, and it is possible to obtain a more uniform output level for varying inputs, since the delay action reduces the possible range of variation of the latter; the greater the delay the more nearly perfect will be the A.G.C. system.

38. **Amplified Delayed A.G.C.—Quiescent A.G.C.**—In general, if a reduction is made in the R/F amplification preceding a detector stage, the peak value of the signal input will be too low for the production of adequate A.G.C. negative bias. In certain cases, however, it is desirable to economise in R/F amplification, in which case the detector may be supplied with signals of the order of 1 volt peak value; the bias voltages arising from rectification must then be amplified before being fed back to earlier R/F stages.

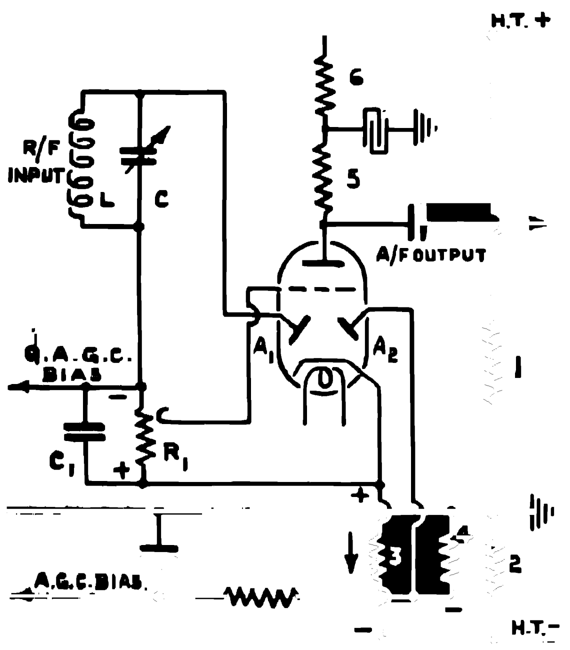


FIG. 30.

Fig. 30 provides the circuit details of an arrangement using a duo-diode-triode. This valve combines the functions of a double diode already discussed and, in addition, the triode section of it is used as an A/F amplifier, and also as a direct current amplifier (K.59).

The signal input is fed into the tuned LC circuit and applied—for detection—between anode A_1 and cathode by means of the ordinary diode circuit. Condenser C_1 is the R/F by-pass, and resistance R_1 is the load resistance across which the small direct current P.D. is developed.

If the incoming signal be a modulated one, A/F variations in potential may be picked off across R_1, the latter being used as a manual volume control, and applied between grid and cathode of the triode portion of the valve. In this case, the circuit has been slightly simplified by omitting the R/F filter circuit which is usually interposed in the A/F feed. The output is shown to be resistance/capacity coupled to a succeeding A/F amplifying stage, the usual decoupling arrangements also being included.

Resistances (1) and (2) together form a potentiometer across the H.T. supply, the general earth line being connected to a point down this potentiometer. The cathode is connected to the H.T. negative terminal through resistance (3);

when no current is flowing through the valve, the effect of the position of the earth line is to make the cathode negative to earth by the value of the P.D. across resistance (2), usually about 100 volts. Since the second anode A_2 is connected to earth through resistance (4), under the above conditions it would therefore be positive to its cathode by a similar amount—namely, about 100 volts. Clearly, current would then flow in resistance (4), and a direct current P.D. would be developed across it. The relative potential of the cathode is, however, not determined only by the P.D. across resistance (2); it is largely controlled by the current flowing through the valve.

Resistances (5), (6) and (3) represent the output impedance of the triode section of the valve. When there is no signal input, the current through the valve will have its maximum value, and it is arranged that the fall in P.D. across resistance (3), produced by the returning space current, is sufficient to make the cathode positive with respect to the earth line by an amount which must depend upon the amplification produced by the valve, usually from 20 to 40 volts positive. The behaviour of the circuit is as if two sources of E.M.F. were connected in opposition, each trying to produce contrary P.D.'s between anode A_2 and the cathode: when the fall in P.D. down (3) is greater than that down (2), the cathode is positive to earth, and under those conditions no current flows in the diode section A_2, and no bias potential is developed across resistance (4).

Let us assume that the cathode is 20 volts positive to earth when the signal input is zero, and that the valve circuit has a V.A.F. of 20. A signal of peak value 1 volt will produce a negative bias across R_1, of about the same value; this is applied between grid and filament of the triode section, the current through the vale is reduced, and the P.D. across resistance (3) falls by about 20 volts. The cathode will then be just at earth potential, effectively the two oppositely connected "batteries" just cancelling each other. Since A_2 is also at earth potential, no current yet flows in that section of the valve.

If the signal increases to 1·5 volts, the P.D. across resistance (3) will fall by another 10 volts, thereby making the cathode 10 volts negative with respect to earth; current then flows in resistance (4) and a negative A.G.C. bias is developed which may be fed back to preceding valves in the usual way. It is therefore apparent that the circuit provides a delay in the application of A.G.C. until the signal input exceeds 1 volt; for greater inputs the requisite bias voltages are provided, the maximum negative bias being limited by the P.D. across resistance (2).

The above circuit may also be used in the provision of Q.A.V.C. One simple way of doing this is to arrange to damp the tuned circuit of one of the R/F amplifying stages, by causing grid current to flow when no strong carrier wave is being received; the stage gain may thereby be reduced, and the set rendered less sensitive and quieter in operation.

In the case of Fig. 30, when no signals (or only weak ones) are being received, there will be little or no D.C. potential difference available across R_1, and the current through the valve will be large. The cathode will therefore be positive to earth, and this positive potential may be fed back to the grid of one of the previous R/F amplifying stages by means of the line labelled "Q.A.G.C. bias." When the grid of that previous stage receives this positive bias, grid current will flow, thereby highly damping the tuned circuit which is attached to it.

When strong signals are received, a larger bias potential appears across R_1, thereby reducing the current through the valve; the cathode ceases to be positive with respect to earth, grid current no longer flows in a previous R/F amplifying stage, and the tuned circuit of the latter is undamped. The receiver quickly returns to its condition of high sensitivity and the station is strongly received; powerful stations come in somewhat suddenly, and the tuning may be described as "ploppy."

39. **Voice Operated A.G.C. Circuits.**—Under ordinary conditions the input energy to a microphone varies over a wide range, and it has been estimated that the power ratio of the strongest to the weakest significant sounds in speech may amount to 10 million to one, or 70 db (Appendix A). It is sometimes considered that this range consists of two components, the "volume range," and the range due to intrinsic differences in the various speech sounds. The volume range of about 40 db. is due to the differences in the speech powers of different announcers, and the way they talk into the microphone; these volume differences can be reduced to a considerable extent by manually

operated gain controls. The remaining variation of about 30 db. is due to the natural differences in energy associated with different speech sounds, including modifications introduced by variations in emphasis. The total power range is so large that some form of A.G.C. is necessary in order that the intensity of the strong vowel sounds may not overload the amplifiers, so producing distortion, and that the output may be maintained approximately constant and independent of the input. This need for a voice operated A.G.C. system applies equally to public address systems and to long distance telephony.

In the case of ordinary R/T receivers, the A.G.C. bias is provided by the carrier frequency wave. In the case of A/F public address circuits, no carrier is available and the gain of the set must therefore be controlled by the speech frequency voltages.

In the case of ordinary A.G.C. circuits, the sensitivity of the receiver is made to depend on the value of the carrier wave amplitude. In the case of voice operated circuits, it is similarly possible to produce an A.G.C. negative bias by rectifying the A/F voltages, and the mean value of this bias may be made to depend upon the *peak values* of the speech frequencies. If the mean power level of speech is high, the peaks of voltage will be correspondingly high, so giving rise to a proportionately larger A.G.C. bias.

Voice operated A.G.C. circuits therefore involve two important requirements :—

(*a*) The action of the A.G.C. in compensating for excessive speech voltage must be as rapid as possible In the condition just before an announcer commences to talk, the A.G.C. is inoperative and the amplifiers are in the condition of maximum sensitivity. As soon as the operator commences to speak, provided his voice is above a certain minimum intensity, he will overload the power amplifiers with consequent distortion, until such time as the A.G.C. has reduced the sensitivity of the set to the required amount. It is therefore necessary that the first few cycles of speech frequency must bring the A.G.C. into operation.

(*b*) The action of the A.G.C. should be slow in the reverse direction when called upon to compensate for a reduction in the intensity of speech. It is clearly desirable to keep the sensitivity of the set adjusted as nearly as possible to the average intensity of the announcer's speech, taken over a period of time ; this will be the case if a large time constant (CR value) is arranged for the condenser/resistance combination in the rectifying circuit handling the A/F voltages.

If these two requirements can be satisfied, the sensitivity of the A/F amplifiers may be kept adjusted by the peak voltages of the speech input, but any sudden *reduction* in speech intensity would not be compensated immediately. The latter condition would only arise in practice in the case of a particularly bad announcer who clipped his words at the ends of sentences, or syllables at the ends of words.

40. **Voice Operated A.G.C. in a Service Type Public Address (Wa/T) System.**—Fig. 31 shows, diagrammatically, the A.G.C. circuits of a particular Service type of input amplifier.

The first valve (1) is a variable mu pentode, the amplification of which is controlled by means of a variable negative bias fed back from the A.G.C. system. The output may be considered to be resistance-capacity-coupled to the second amplifying stage (2) ; it is proposed to treat at a later stage the principles of the tone control circuits which add complexity to the output impedances of valves (1) and (2). For the purpose of the present argument, therefore, the essential part of the output of valve (2) will be taken to be resistances (3) and (4). The A/F variations in potential which are available across these output impedances are used to feed the subsequent intermediate amplifier (input voltage $\mathcal{V}_1$), and also to supply the signal input between grid and filament of the A.G.C. amplifying valve (5). This valve is transformer coupled to an ordinary diode detector (6). The circuit is a conventional one, condenser (7) is an A/F by-pass condenser, and therefore has the large value of 1 µF., and resistance (8) is the load resistance of value 1 megohm, across which the rectified D.C. potential appears. The value of this D.C. potential is proportional to the magnitude of the A/F input ($\mathcal{V}_2$) to valve (5). The circuit shows that the available negative bias is fed back to

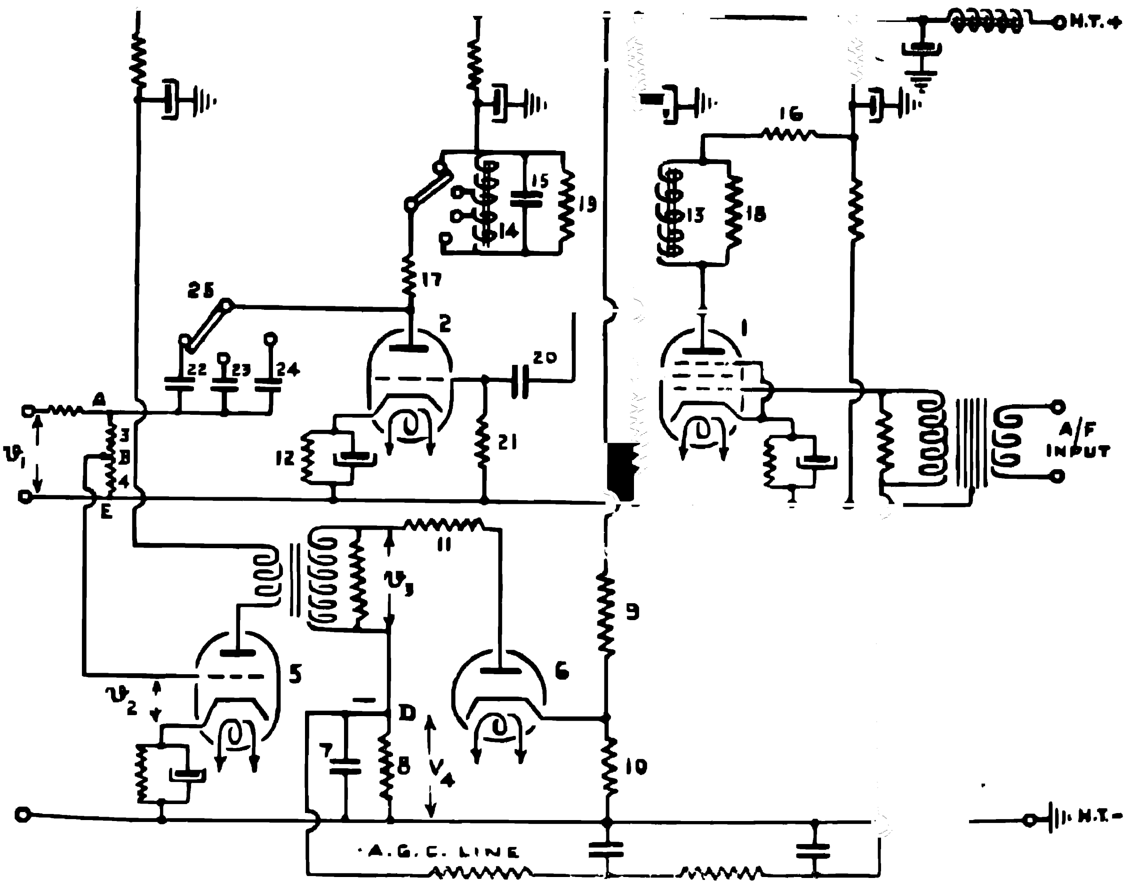


FIG. 31.

valve (1) through the usual "low pass" filter. The filter is designed to pass the relatively slow changes of rectified potential (V_4) which are required to control the amplification of valve (1), but to obstruct any A/F variations which would otherwise be transferred from valve (5) back to the grid of valve (1); any such transfer would result in a "howl" and produce considerable distortion.

The diode is arranged so as to produce amplified delayed A.G.C., and the A.G.C. bias does not come into operation until the peak value of the voltage $\mathcal{V}_3$ exceeds the "delay voltage." This delay voltage is arranged by means of resistances (9) and (10) which form a potentiometer across the H.T. supply, the cathode of valve (6) being connected to an intermediate point. In practice, the cathode is maintained at a potential of about 50 volts positive to earth, but the value can be adjusted by the size of resistance (10). Accordingly, no rectification takes place in valve (6) until the peak value of the signal input is at least equal to the delay voltage. When the peak value of the transformer secondary voltage ($\mathcal{V}_3$) exceeds the delay voltage, the anode of the diode will become positive with respect to the cathode, and a current will flow through the valve and the resistance (8) during the peaks of the positive half cycles.

Condenser (7) becomes charged very *quickly*, the time occupying a small fraction of a second; the charging rate is controlled by resistance (11). The diode valve continues to rectify until the delay voltages plus the rectified voltage V_4 are together approximately equal to the peak value of the voltage $\mathcal{V}_3$. If $\mathcal{V}_3$ remains steady, the rectified potential V_4 will also remain steady, the valve passing sufficient current during the positive half cycle peaks to compensate for that which leaks *slowly* away through the high resistance (8). If the voltage input $\mathcal{V}_3$ falls, rectification in the diode (6) will cease, and the charge in the condenser will leak *slowly* away through the high resistance

until the delay voltage plus the rectified voltage is again equal approximately to the peak value of the signal input, or until the rectified voltage is zero, if $\mathcal{V}_3$ is less than the delay voltage. It may be noted that using the quoted values of condenser (7) and resistance (8), the CR value of the combination is one second (*cf.* paragraph 27).

The whole action of the circuit may be summarised as follows :—

(*a*) When the A/F input is below a critical value, determined by the delay voltage, no A.G.C. action occurs and the circuit is maintained in the condition of maximum sensitivity.

(*b*) When the A/F input increases above that value, rectification occurs in valve (6), the condenser (7) charges up rapidly, negative grid bias is applied rapidly to the grid of valve (1), and the amplification is quickly reduced to compensate *almost* completely for the increased A/F input. As in previous cases, compensation cannot be complete, because some increase in the A/F input to valve (5) is necessary in order to produce the compensation.

(*c*) If the A/F input falls, $\mathcal{V}_3$ will decrease, condenser (7) will slowly discharge, and thereby slowly increase the amplification produced by valve (1). In the course of normal speech, the requisite degree of compensation will be determined by the highest peaks of A/F voltage.

The relative values of resistances (3) and (4) determine the ratio of $\mathcal{V}_1$ to $\mathcal{V}_2$. The adjustment is set so that $\mathcal{V}_1$ is sufficient fully to load the subsequent power amplifiers when the A.G.C. is in operation.

41. **Automatic Grid Bias.**—In Fig. 31, automatic grid bias is provided independently for each valve except the rectifier (6). The principle is that the grid of each valve is anchored to earth potential, and the cathode is raised to a potential positive with respect to earth by the insertion of a resistance between the cathode and earth. The total anode current (or space current) will then pass through that resistance, making the cathode positive with respect to the grid ; this is equivalent to making the grid negative with respect to the cathode, *i.e.*, negative grid bias is applied The magnitude of the bias is given by the product of the anode current and the resistance employed , for example, in the case of valve (2), the resistance (12) is 600 ohms and the anode current is of the order of 3·8 milliamps, producing a negative bias of about 23 volts.

It is necessary that the voltage developed across a bias resistance should depend on the mean anode current, and should not follow any changes in the latter taking place at audio-frequencies. For this reason, a condenser of high capacity, usually an electrolytic condenser, of value about 20 microfarads, is shunted across the automatic grid bias resistance in each case, to by-pass to earth the audio-frequency components of the anode current. One advantage of automatic grid bias obtained by this means is that it is, to a large extent, self-adjusting. If a valve is used which passes more anode current than normal for a given grid bias, the additional anode current will provide additional negative grid bias. This in turn will reduce the anode current of the valve, and an equilibrium point will be reached at which the increase in anode current is much less than the amount it would be if a fixed grid bias were employed. A similar but opposite effect occurs if a valve is employed which passes less anode current than normal for a given value of grid bias. Furthermore, if the high tension supply voltage varies, the action of the anode current of each valve produces corresponding and compensating changes in the value of the grid bias.

The provision of automatic grid bias in this way is not restricted to indirectly heated valves, although with them it is simplest in operation. In the case of directly heated valves, the situation is complicated by the fact that the two ends of the filament are not at the same potential. Fig. 32 illustrates the use of automatic grid bias in the case of two triodes in push-pull in an A/F amplifying stage, the filaments being directly heated by the A.C from two separate transformer windings. Each winding is centre tapped and connected to earth and H.T. negative by resistances (1) and (2) respectively. The anode current flows through these resistances, and, if the stages are perfectly balanced, raises the filaments to equal positive potentials above earth.

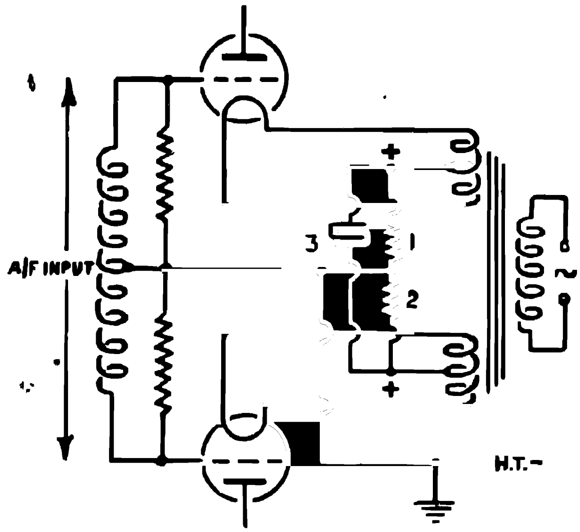


FIG. 32

It may be noted that with this push-pull arrangement, there is no need to provide an A/F by-pass condenser between the centre tap of the input inductance and the centre tap on the two transformer windings; with a symmetrical input the centre tap of the input inductance should always be at an unvarying A/F potential, and, for this reason, it is only necessary to arrange that the centre taps of the filament windings are themselves always at the same A/F potential. This is effected by means of condenser (3).

42. **Tone Control.**—Tone *correction* implies the adjustment of the relative strength of reproduction of different audio-frequencies. This is usually effected by altering the frequency response characteristic of the amplifier system, by the use of suitable networks involving capacity, inductance and resistance.

There are numerous causes giving rise to the necessity for tone control. For example, side-band cutting may make it advisable to emphasise the high audio-frequencies with relation to the lower ones. On the other hand, the necessity may arise entirely due to peculiar acoustic conditions; in paragraph 6, it is observed that the poor acoustic conditions which exist in a ship make it necessary to attenuate audio-frequencies below 800 cycles/second, in order to reduce " boom " and " echo," and to accentuate frequencies above 2,000 cycles/second to improve the intelligibility of speech.

There are four principal forms of tone correction, namely the accentuation or attenuation of high or low notes respectively; the special needle scratch filters used in gramophone pick-ups form a special case of tone correction.

High note accentuation will be required when the consonants of speech (*i.e.*, letters S, T, F, etc.) are weak, music is too mellow, and the reproduction is " woolly."

Its introduction will improve intelligibility.

High note attenuation will be required when the consonants of speech are harsh and over emphasised (whistling S's, etc.), and the reproduction generally is hard and shrill, and too " brilliant." High note attentuators will reduce gramophone needle scratch, and reduce the " mush " in radio reception.

Low note accentuation will be required when music is lacking in " body." This type of correction is only of value when loudspeakers and amplifiers capable of reproducing very low notes are used; in other cases there is a risk of overloading the amplifier and damaging the loudspeaker. **Low note attenuation** is usually only necessary in cases where it is desired to eliminate " boom," possibly due to bad acoustic conditions.

43. **General Principles of Tone Corrector Circuits.**—In tone corrector circuits, use is made of a combination of some of the following principles :—

(*a*) That the impedance of a pure resistance is independent of the frequency.
(*b*) That the reactance of an inductance varies directly with the frequency.
(*c*) That the reactance of a condenser varies inversely with the frequency.
(*d*) That an acceptor or rejector circuit has a minimum or maximum impedance respectively at a certain frequency.

In general, a suitable network is constructed so that the impedance varies in a manner corresponding with the required variation of response or amplification. An *uncorrected E.M.F.* is then applied to this circuit through a resistance (which may be a valve), which is larger than the highest value of the impedance of the tone corrector network within the required A/F range. The former resistance, being the largest impedance in the circuit, then controls the current through the whole circuit, which is hence directly proportional to the uncorrected E.M.F. at all audio-frequencies within the required limits. The E.M.F. developed across the tone corrector network at any frequency will then be given by the product of the current and impedance at that frequency; the P.D. developed across the corrector network will then be an *E.M.F. which has been corrected for tone*, and it may be passed on to subsequent A/F amplifying stages in the ordinary manner.

In most cases of high or low note accentuation, the accentuation is only relative. In high note accentuation, all frequencies except the high notes are attenuated, thus leaving the high notes relatively accentuated, and similarly with low note accentuation. Most common forms of tone correction reduce the general volume level by an amount which must be compensated for by increasing the response to all frequencies uniformly, conveniently done by means of one of the gain controls in the amplifier system. Certain simple tone controls are well known. A resistance shunted across the terminals of a gramophone pick-up will produce attenuation of the higher audio-frequencies; the lower the resistance the more drastic the effect. In radio receivers, strong use of reaction sharpens the tuning and produces sideband cutting. In superheterodyne receivers, with variable band pass coupling for the I/F transformers, the higher audio-frequencies are strongly attenuated by employing a loose coupling.

It is proposed to illustrate the principles of tone correction by reference to one particular case

44. **Tone Correction in a Particular Case.**—The amplifier circuit of Fig. 31 provides an example of high note accentuation and low note attenuation.

HIGH NOTE ACCENTUATION CIRCUITS.—Two separate high note accentuation circuits are provided. A fixed degree of accentuation is given by inductance (13), and a variable amount by the tapped inductance (14) assisted by the condenser (15).

In each case the principle is the same. At all frequencies below about 1,000 cycles the impedance of the choke (13), and of the LC circuit (14) and (15), is small compared with the coupling resistances (16) and (17) respectively. At these frequencies, therefore, the coupling is effectively pure resistance-capacity, and is approximately equal at all frequencies, except for the action of the low note attenuation circuits described below. The resistances (16) and (17) are, however, much lower than the optimum value necessary for full amplification. At frequencies above about 1,000 cycles per second, the impedances of the inductance (13) and the LC circuit (14) and (15) become greater than the resistances (16) and (17) respectively, thereby increasing the amplification at those frequencies. Furthermore, in each case at all frequencies above the critical frequency where the impedance of the inductance, or resonant circuit, is equal to the value of the coupling resistance, the impedance of the whole anode circuit will vary directly with the frequency, thereby providing a rising frequency response characteristic (paragraph 43). At very high frequencies, excessive rise in the impedances of the inductance (13), and the LC circuit (14) and (15), is limited by the resistances (18) and (19).

The condenser (15) in conjunction with the inductance (14) produces a resonant circuit providing a maximum impedance at about 6,000 cycles. The resistance (19) prevents this resonance from being too sharp, but permits the resonant circuit to provide greater accentuation of frequencies in the neighbourhood of 3,000 to 6,000 cycles per second than is provided by the plain inductance (13).

The variable inductance (14) is the high note accentuation control. In the upper position of the switch the tone control circuit is inoperative, and in the lower position it exerts its maximum effect.

LOW NOTE ATTENUATION CIRCUITS.—Two separate low note attenuation circuits are provided. A fixed degree of attenuation of frequencies below about 400 cycles per second is

provided by the condenser (20) and resistance (21). A variable amount of attenuation of frequencies below about 800 cycles per second is provided by the variable capacity (22) to (24), acting in conjunction with the total resistance between the points A and E (100,000 ohms). The principle is the same in both cases.

In the case of the fixed low note attenuation circuit, attenuation starts at the frequency when the reactance of the condenser (20) is equal to the resistance (21). As the frequency falls, the reactance of the condenser ($0 \cdot 002$ μF.) rises, and this reactance acts in conjunction with the resistance (90,000 Ω) virtually to form a potentiometer which supplies to the grid/filament circuit of valve (2) only a proportion of the total voltage developed between anode and filament of the valve (1). The lower the frequency the less will be this proportion, and the relative intensities of the low notes are thereby reduced.

The action of the variable low note attenuation circuit is similar, except that the critical frequency at which low note attenuation commences is adjusted by varying the capacity (22) to (24), by means of the low note attenuation control switch (25). When this control is over to the left of the diagram, the capacity is large ($0 \cdot 1$ μF.) and the attenuation of the low notes is negligible in this circuit, and the overall low note attenuation consists only of that provided by the condenser (20) and resistance (21). When the low note attenuation control is over to the right, this circuit attenuates all frequencies below about 800 cycles per second [condenser (24) is $0 \cdot 0002$ μF.]. This effect is additive to the fixed attenuation of frequencies below about 400 cycles per second, provided by the fixed attenuation circuit, and the overall effect is, therefore, a *moderate* attenuation of frequencies between 800 and 400 cycles, and a *severe* attenuation of frequencies below 400 cycles per second.

45 **A/F Amplification.**—Owing to the difference in the frequencies concerned, there is an essential distinction in technique between R/F and A/F amplification, a distinction which becomes accentuated in the design of R/T receivers where little distortion can be permitted.

With the exception of the modifications rendered necessary by tone controls, the general requirement in A/F amplifying stages is that the voltage developed across the output circuit should be a replica on a larger scale of the input voltage applied between grid and filament. In order to avoid distortion (paragraph 7), the V.A.F. of the stage should be independent both of the frequency and the amplitude of the input voltage over the required range of frequencies and amplitudes.

Using triode valves, **amplitude distortion** can arise from various causes, one of the commonest being the application of too large an input voltage. The grid swing is necessarily larger in A/F than in R/F stages, and it is a matter of some difficulty to get a valve dynamic characteristic which is straight over a large enough range of grid voltages. The higher the ratio of the external anode impedance to the valve A.C. resistance, the more likely is this condition to be satisfied. In addition, any flow of grid current during the positive half cycles will damp the input circuit and cause amplitude distortion ; high anode voltages are necessary to obtain a sufficiently large amount of the straight part of the characteristic with respect to the working point employed. The advantages of using valves in various forms of push-pull are explained in Section " F." One of the most popular forms of A/F output stage consists in a curvature biased **class " B " (Q.P.P.) push-pull** arrangement. The two valves in push-pull are commonly enclosed within the same glass envelope, a characteristic feature being that the combination is worked with little or no negative bias. Grid current flows during each half cycle, but distortion is cancelled by the push-pull arrangement, as explained in Section " F." The input impedance of a class " B " push-pull combination of this nature is necessarily low, and it is usually matched to the output of the preceding A/F amplifying stage (the driver valve), by means of a step-up transformer, the circuit being similar to that shown in Fig. 34. By removing the usual disadvantage associated with the flow of grid current, considerable output power can be obtained from a class " B " output stage.

In the case of transformer coupled A/F amplifying stages, a common source of distortion is due to the approach to saturation of the iron core produced by too large a mean value of the anode current. This can be considerably reduced by feeding the H.T. to the anode through a separate

choke in parallel (parafeeding) ; the choke carries the steady anode current, and the arrangement is similar to a parallel fed transmitter (K.24). The principle is illustrated in Fig. 34, stage 1.

The presence of amplitude distortion may be detected by means of the harmonics thereby generated ; the tone quality is affected very considerably. It is important, however, to realise that not all of the amplitude distortion arises in the post detector stages ; much of it may be due to the use of a non-linear detector, or to distortion of the amplitude envelope of the incoming R/F signal in the R/F amplifying stages.

Frequency distortion will be produced unless the output impedance is independent of frequency. It is therefore evident that tone controls deliberately introduce frequency distortion. Some of the older Service A/F amplifiers were noted for the peaky amplification produced at certain frequencies—about 2,000 cycles per second.

From the point of view of R/T reception, the best intervalve coupling would therefore be resistance-capacity coupling. In Section " F " it is observed that this method is not very suitable for radio-frequencies, owing to the effects of self-capacity ; that effect is less pronounced at audio-frequencies, but it is necessary that the reactance of that capacity should be high at the highest audio-frequencies dealt with, in order to avoid attenuation of those frequencies. This requirement limits the size of the resistance employed. Similarly, the coupling condenser must be large enough for its reactance to be negligible compared with the resistance of the grid leak, in order to avoid attenuation of the lower frequencies. The time constant of the grid coupling condenser and leak must be kept as small as possible, consistent with the foregoing requirement, otherwise overloading of the grid, and the consequent flow of grid current, may increase the negative grid bias so much that the grid swing reaches the bottom bend of the characteristic and amplitude distortion is produced.

In the Service, typical values for the coupling condenser range from 0·01 μF. to 0·1 μF., using about 2 megohms for the value of the grid leak. These values may be compared with those, used in R/F amplifiers where the value of the condenser ranges from 0·001 μF. to 0·0001 μF. at high radio-frequencies, using a grid leak of 1 megohm. In commercial broadcast receivers, the common value of the coupling condenser in an A/F stage is 0·1 μF., working with a grid leak of 0·5 megohms or less.

The above limitation in the size of the output resistance requires that a valve of correspondingly low A.C. resistance be used, if a reasonably straight dynamic characteristic is to be obtained, and amplitude distortion avoided. A common practical figure for the external resistance is 25,000 ohms, and, with this value, using a triode valve, the valve A.C. resistance should not exceed 7,000 ohms ; this combination will give nearly uniform amplification from 30 to 10,000 cycles per second. Using a pentode output valve, the typical value of the output impedance is somewhat lower.

Transformer coupling is also often employed in A/F stages, particularly with the class " B " push-pull arrangement to which reference has been made ; in the latter case, however, the output transformer is a step-down one in the output direction. In general, care must be taken that the transformer windings do not tune with their (self or stray) capacities anywhere near the audible range, otherwise frequency distortion will be produced.

To give a V.A.F. substantially independent of frequency, the reactance of the primary of the step-up transformer at the lowest audio-frequency must be large compared with the A.C. resistance of the valve. It follows, therefore, that for a given valve and transformer, low audio-frequencies below a certain frequency will not be satisfactorily amplified. Common values of the primary inductance range from about 20 henries to 100 henries, the bulk of the transformer increasing very much with the inductance. Since the dimensions of the output stage are limited, this leads to the use of transformers with only a small step-up ratio. In addition, the smaller the secondary, the less is the effect of its self-capacity in producing attenuation of the higher audio-frequencies.

46. **The Output Stage.**—In a receiver having several stages of A/F amplification, the earlier ones aim chiefly at voltage amplification, but the last one must be designed to deliver as much power as possible to the output impedance—usually a loudspeaker system.

In the case of a triode, it has been seen (in Section " F ") that the choice of the output impedance necessary for the production of *maximum power*, is inevitably accompanied by considerable distortion, produced by the flow of grid current. In the case of a pentode output, as in the case of the triode, a compromise has to be found between the conflicting requirements of distortionless reproduction and large power output. In practice, the ear can tolerate a certain amount of distortion, and a second harmonic content up to about 5 per cent. is usually allowed. The valve maker arrives at an estimate of the optimum load impedance to be used with any A/F amplifying valve, by considering jointly these two associated requirements.

In the case of a pentode, the output impedance giving the maximum power output cannot be shown to be equal to the A.C. resistance of the valve.

Among the common forms of output stage are included push-pull arrangements, employing triodes or pentodes in their class " A " or class " B " forms. A recent modification is sometimes used, and is found bearing the name " low loading push-pull " or " class A-B push-pull " amplifier. Power triodes are still used, but the output pentode is generally preferred where a high power output must be produced as economically as possible.

It is proposed to refer in slightly greater detail to the output pentode, and the class " A-B " amplifier.

47. **The Output Pentode. Power Output.**—Fig 33 represents the anode characteristics of a typical output pentode, capable of a maximum anode dissipation of 10 watts using 250 volts on the anode. Inspection of the characteristics suggests that with 250 volts on the anode a suitable working point might be obtained with the grid biased to - 15 volts, to the point O. The performance of the valve when operating under dynamic conditions can now be estimated by drawing load lines passing through the operating point O. When the load is small the load line is steep, and the figure shows EF the load line for 5,360 ohms; large grid swings produce large alterations of anode current but small alterations of potential across the load. As the load increases the load line becomes less steep, BOD being the load line for the large load of 31,250 ohms.

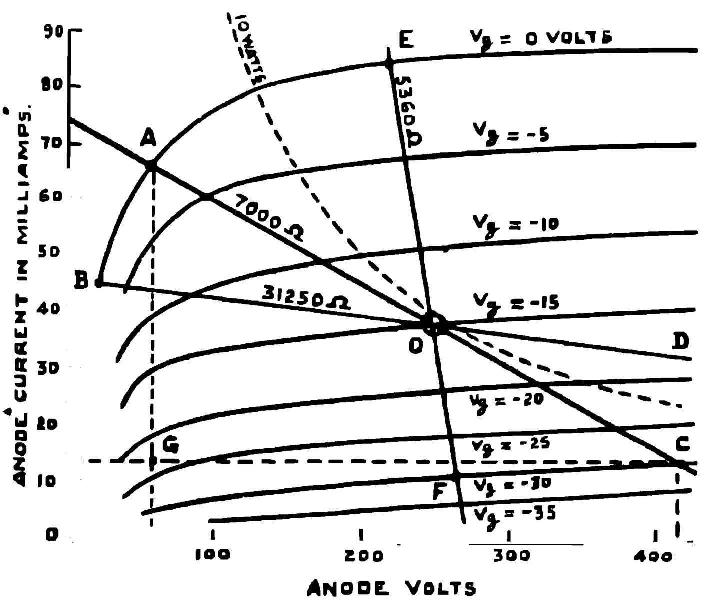


FIG. 33

Dealing with the case of a triode, it was appreciated in Section " F " that **distortionless amplification** involves a load line which makes equal intercepts with a series of equally spaced anode characteristics. In this case, if the grid is assumed to swing with an amplitude of 15 volts about the point O, its excursion will take it to the point A and the points C respectively on the load line AOC. The 15 volt swing from O to A drops the anode volts from 250 to about 56, a difference of 194 volts; a similar swing from O to C increases the P.D. between anode and filament to about 412 volts, an increase of 162 volts. It is clear, therefore, that the A/F output voltage is not a replica of the input grid voltage, since equal

changes of grid voltage in opposite directions do not produce equal changes of anode voltage. Distortion is present, but a rough practical rule states that the second harmonic content will not exceed 5 per cent. if the ratio of OC to OA is about 9 : 11. As the load decreases, the distortion usually becomes less, but the power output falls. When the load is very high, as in the case of the BOD load line, the distortion is very considerable. Moreover, the very rapid rise in anode voltage when the grid swings from O towards D, constitutes a source of danger ; voltages of over 1,000 frequently develop and are liable to cause a breakdown in the valve or insulation, and the situation would be aggravated if, for example, the loudspeaker were disconnected while the valve is operating. A compromise has therefore to be made, and is represented in this case by the load line AOC ; it provides a high power output combined with reasonable freedom from distortion.

The mean power output may be determined from the peak values of the anode current and anode voltage swing. With a 15 volt grid swing, in Fig. 33, the anode current varies from A to G, the anode volts similarly varying from G to C. If the amplitude of the anode current is given by $\mathscr{I}_a$ and that of the anode volts is given by $\mathscr{V}_a$, we have $\mathscr{I}_a = AG/2$ and $\mathscr{V}_a = GC/2$. The power output is given by the product of the R.M.S. current and the R.M.S. volts, hence

$$\text{Power output} = \frac{\mathscr{I}_a}{\sqrt{2}} \times \frac{\mathscr{V}_a}{\sqrt{2}}$$

$$= \frac{AG \times GC}{8} \quad \ldots \text{ in watts if } \mathscr{I}_a \text{ and } \mathscr{V}_a \text{ are in amps and volts respectively.}$$

Taking numerical values from Fig. 33—

For load line AOC ...... Power output $= \dfrac{0{\cdot}052 \times 356}{8} = 1{\cdot}7$ watts.

For load line EOF ...... Power output $= \dfrac{0{\cdot}074 \times 45}{8} = 0{\cdot}4$ watt.

Power output under static conditions at O $= 0{\cdot}038 \times 250 = 9{\cdot}5$ watts. (Note the 10 watt **static output curve** of Fig. 33.)

The output pentode offers the advantage of providing high amplification combined with economy in valve stages. A disadvantage is that the tone quality frequently leaves much to be desired. The amount of distortion is found to depend upon the size of the output load. The third harmonic distortion increases steadily with the load, but the second harmonic distortion shows a minimum value for a certain critical value of load ; it is usually that value of the output load which is used by the valve makers in arriving at a figure for the *optimum output impedance.* Where the output from the pentode is applied directly to a loudspeaker, the requisite impedance matching is usually effected by means of a transformer coupling. For example, a 13 ohm loudspeaker could be matched with an 8,000 ohm output impedance by using a transformer with step-up ratio (T) roughly 25 : 1, since $8000/T^2 \doteqdot 13$.

48. **Low Loading Push-pull, or Class "A-B" Push-Pull.**—The system of push-pull amplification referred to by this name, is of a type specially developed to produce a large undistorted output using small valves with low maximum anode dissipation. The title "low loading" is used because the principle employed requires that, for maximum power output, the anode to anode load should be less than that employed in normal class "A" amplification. The reason for the term "class A-B" is that the system provides a compromise between the principles of class "A" and class "B" amplification. (*Cf.* F. 31.)

By employing an anode to anode load which is lower than normal, and only slightly above the total internal impedance of the output valves (*i.e.*, twice the impedance of each valve considered singly), the percentage distortion is decreased. By operating the valves lower down the anode current/grid volts characteristic, the efficiency is slightly increased. The result of these two operations is a reduction in the third harmonic content, but an increase of the second harmonic

one. If the two halves of the push-pull stage are accurately balanced, the second harmonic distortion will cancel out in the windings of the output transformer. It is often considered that the tone quality obtainable with a push-pull triode combination worked under class "A-B" conditions, is superior to that obtainable under class "B" conditions.

49. **Typical Straight Set R/T Receiver.**—Fig. 34 gives the circuit details of a receiver employing two R/F amplification stages using screen grid valves, followed by an indirectly heated valve power grid detection stage, the output stage consisting of two neutralised triodes in push-pull.

In the aerial circuit, a small condenser (1), of value 0·0001 μF, is present to minimise the effect of variations of aerial capacity upon the tuning. The variable selectivity band-pass filter (2) enables the frequency response characteristic to be adjusted to suit the frequency. The high resistance potentiometer (3) controls the input to the R/F stage and constitutes a manual volume control. The coupling condenser to the first stage is necessary to prevent the grid bias battery from discharging through the tuning inductance. The output impedance to the first R/F stage is a tuned anode output circuit (8), the anode of the valve being parallel fed from resistance (5) and R/F choke (4), the process involving the use of H.T. blocking condenser (7). The grid input circuit to the second stage (8) is tuned, and achieves a double object. The impedance at audio-frequencies of this circuit is negligible compared with the reactance of condenser (7), and so practically no audio-frequency P.D. can be developed across the input to the second stage; "back coupling" is sometimes a cause of the development of A/F potential differences across the input of R/F stages.

Throughout the circuit, resistances and condensers (5) constitute decoupling units. Moreover, resistances (6) form potentiometers across the H.T. supply, and provide the necessary screen voltages.

The choke and by-pass condenser (10) provide a partial separation of the R/F and A/F components, largely eliminating the former one.

Resistance-capacity coupling is used to provide almost uniform amplification at all audio-frequencies. The transformer coupling (12) provides the input to the push-pull stage. The blocking condenser, between resistance (11) and volume control (13), is present to avoid short circuiting the H.T. supply. Volume control (13) controls the input to the primary of the transformer.

In the secondary circuit of the transformer, resistances (14) are present to damp out any parasitic oscillations that may be set up at this stage. They also act as R/F suppressors to any residual R/F currents. The neutralising condensers (15) are sometimes necessary to counter a tendency to self-oscillation. Instability due to feed back of energy through the anode-grid inter-electrode capacity sometimes occurs in the output stage, a fact which points to the advisability of using a screened pentode push-pull stage instead of the more cumbersome neutralised triodes.

The push-pull stage, as represented in the diagram, may be considered to be operated under either class "A" or class "B" conditions; the determining factor is the grid bias. By eliminating the grid bias, and neutralising condensers, by using a valve having the two triodes within one envelope, and by employing a step-down transformer, the output circuit would present the characteristic appearance of a "class B with zero grid bias output" (paragraph 45).

50. **Sound Reproduction.**—This is the process by which the A/F oscillatory currents are made to produce sound waves of corresponding frequency and amplitude. In principle, it involves an instrument which is the converse of a microphone, and the reproducing devices known as **loudspeakers** may be similarly classified into three electrical types, the electro-dynamic, piezo-electric and capacity types. In all cases, A/F oscillatory currents give rise to an alternating mechanical force on a diaphragm, and the to-and-fro motion of the diaphragm produces alternate compression and rarefaction of the air in the vicinity, i.e., sound waves.

The arrangement in a loudspeaker, whereby an audio-frequency alternating current is made to produce an audio-frequency mechanical vibration of a diaphragm, is usually known as **the drive.** A loudspeaker may be regarded as a loudspeaking microphone or telephone

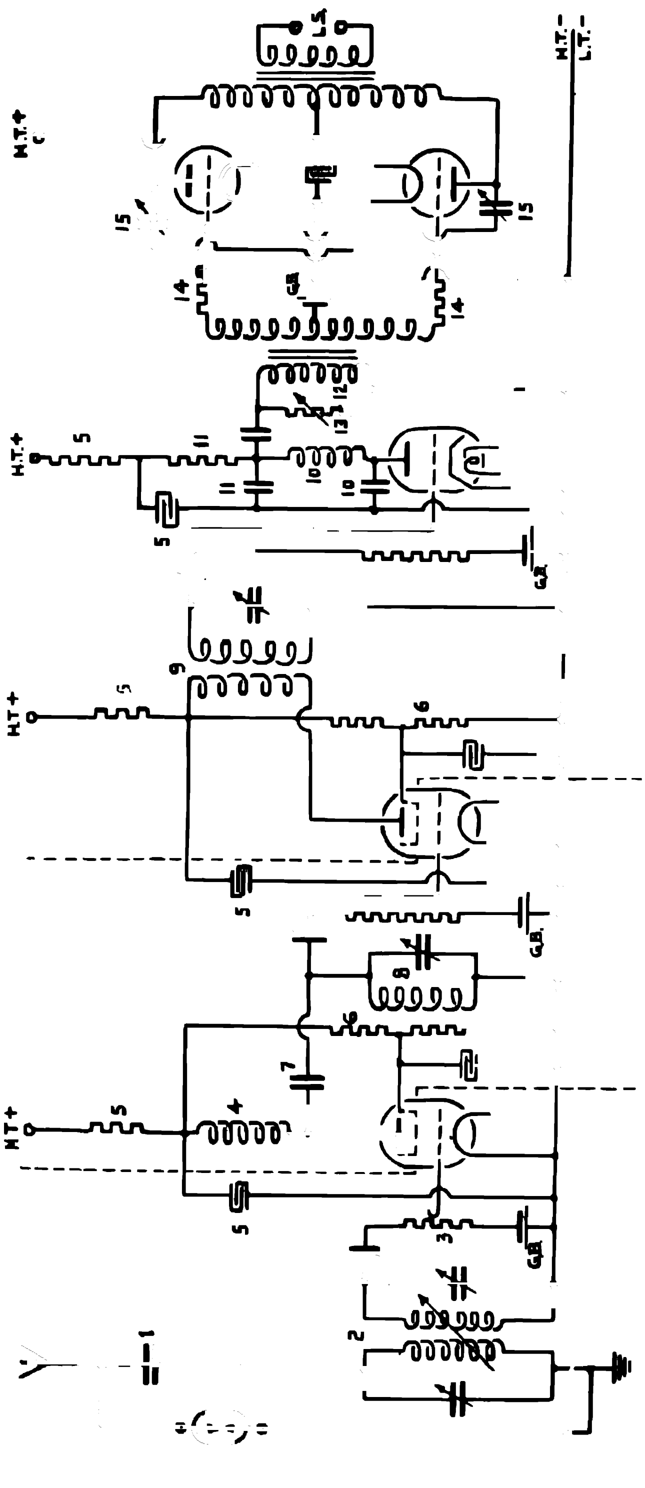


Fig. 34

For perfect reproduction, the intensity of the sound, or the "acoustic output," must bear the same proportion to the electrical power input at all input frequencies and amplitudes ; most loudspeakers have resonances at particular frequencies, very few of them even approaching the condition of distortionless reproduction.

Apart from the electrical division of loudspeakers into three classes, there is also a broader grouping into two main categories—the "horn type," and the "cone diaphragm" type. This grouping is based jointly on the appearance and mechanical functioning of a loudspeaker.

51. **Horn Type Loudspeakers.**—In this type, shown in Fig. 35 (*a*), a large horn is used to provide a big moving column of air so that a *small* diaphragm may be able to produce a large volume of sound. This has the advantage that a correspondingly small electrical input is required to agitate the diaphragm. To produce a plane sound wave the shape of the horn must follow an "exponential" curve, i.e., cross-sections at equal axial intervals from the "throat" of the horn are such that the ratio of the area of any one to the succeeding one is constant. If this is not the case, the sound waves reflected from the inside of the horn produce distortion in the resultant sound issuing from it.

For good quality, a large horn is also essential. The smaller the horn, the higher is the frequency below which the volume of sound from the loudspeaker falls off from the proportionate volume at higher frequencies. This frequency, known as the "lower cut-off frequency," is as high as 300 cycles per second in an ordinary small horn-type loudspeaker. In talking picture installations, horns about 15 ft. long and 4 ft. across at the wide end are used. These bring the lower cut-off frequency down to about 70 cycles per second. To economise in space, a folded exponential horn is often used.

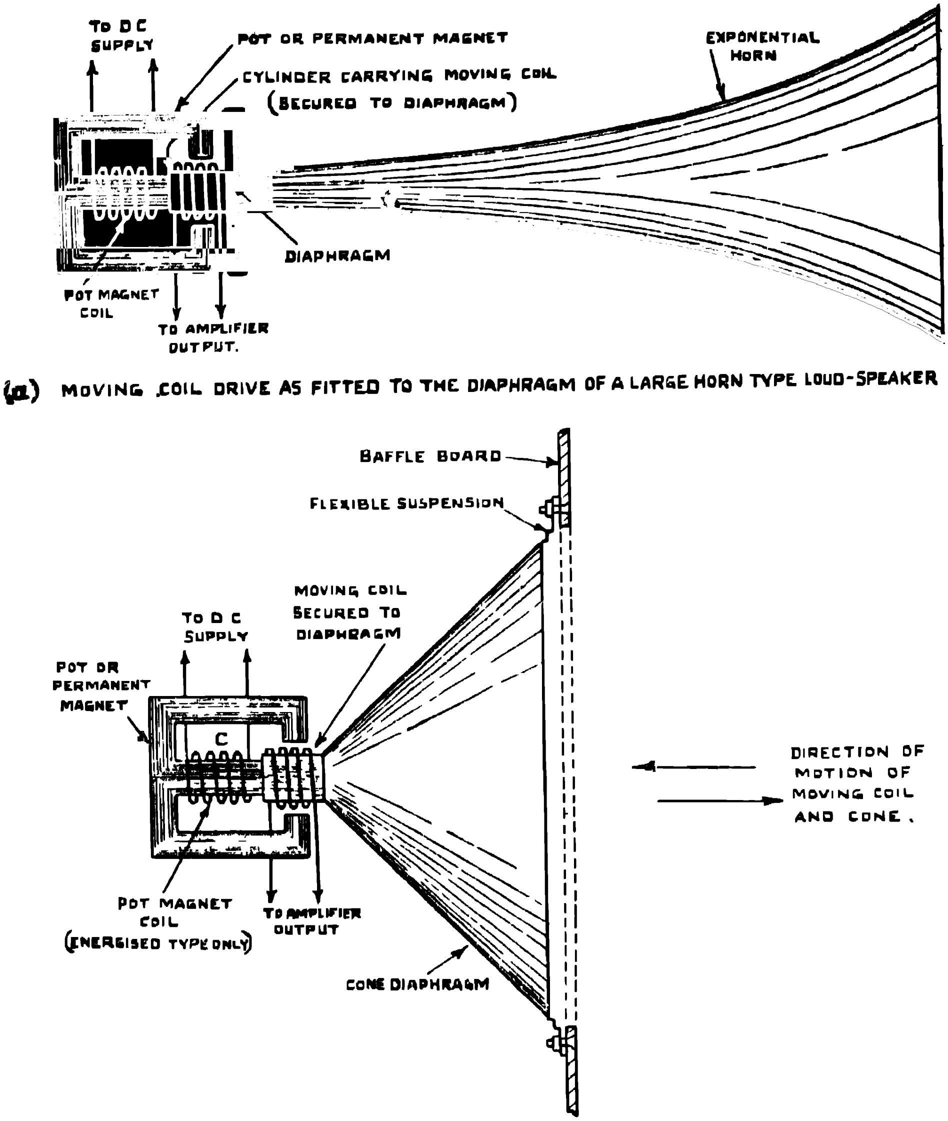
52. **Cone Diaphragm Type Loudspeakers.** The diagram in this type— shown in Fig. 35 (*b*) —is conical in shape and comparatively *large*, so that it acts directly on a large volume of air, no horn being employed. In order to obtain efficient reproduction of low frequencies, it is also necessary to separate the air behind the diaphragm from that in front of it, by means of a large plane board (generally made of wood), called a **baffle** ; otherwise, the air in front of the diaphragm, when compressed, can return to normal pressure by escaping to the back of the diaphragm, instead of compressing the air exterior to itself and so propagating a sound wave. This is obviously most likely to be pronounced at low frequencies, where the time between successive compressions is greatest. In other words, the smaller the baffle, the higher is the lower cut-off frequency.

Diaphragms are usually made from paper, linen, balsa wood, or aluminium alloy. The last is the most efficient.

53. **Loudspeaker Driver Systems.**—The nature of the driving system must depend upon the electrical principle being employed. Where the latter is electro-dynamic, the driving systems are either of the **moving coil** or **moving iron** type, and the principles involved are the same as in the measuring instruments of these types described in Vol. I.

Most of the early types of loudspeaker were of the moving iron type. They are comparatively insensitive, and usually give poor reproduction of low frequencies ; in addition, they have a high impedance varying very considerably with frequency, a fact which makes it impossible to match their impedance to that of the output valve over the requisite range of audio-frequencies. Although these loudspeakers are still considerably used, they have been largely replaced by those of the moving coil type. The latter possess greater sensitivity and a more constant impedance over the A/F range ; it is the only type that will be described in any detail here. In essential details they resemble moving coil microphones (paragraph 11).

54. **Moving Coil Loudspeaker.**—Fig. 35 is a diagrammatic representation of this instrument. A coil of wire wound on a light cylindrical former—the moving coil—is suspended so that it can travel backwards and forwards in the direction of its axis, through a distance of about 0·25



(a) MOVING COIL DRIVE AS FITTED TO THE DIAPHRAGM OF A LARGE HORN TYPE LOUD-SPEAKER

(b) MOVING COIL DRIVE AS FITTED TO A CONICAL PAPER DIAPHRAGM.

FIG. 35.

to 0·5 in., but is not capable of movement in any other direction. It lies in the field of a permanent or electro-magnet, of the shape illustrated in the figure, and called a pot magnet. This field is radial, and therefore at right angles to the coil of wire at every point. The coil, when carrying current, is thus acted on by an electrodynamic force (Vol. I), which sets it in motion at right angles to the plane of its own current and the direction of the field. According to the direction of the current, the coil therefore moves in the direction of its axis to right or left, as determined by Fleming's Left Hand Rule. When an alternating current flows in the coil, the direction of motion is reversed every half-cycle, i.e., the coil undergoes a vibratory motion. The coil is rigidly secured to the diaphragm, and so the latter partakes of the same motion and sets up an air vibration or sound wave of corresponding amplitude and frequency.

The diaphragm and coil are usually suspended by a ring of silk, leather, or other flexible substance round the circumference of the diaphragm. A "spider" in the centre ensures that only axial movement is possible

The chief advantage of the moving coil drive, is the relatively large movement permitted by the free suspension. This has the effect in practice of giving better reproduction at low frequencies than is commonly the case with other types of drive, where the necessities of design lead to a much smaller maximum permissible movement of the diaphragm.

To prevent **frequency distortion,** the impedance of the output valve circuit, including the loudspeaker impedance, should be independent of frequency in the ideal case. The coil, however, inevitably possesses a certain amount of self-inductance. In addition, the movement of the coil in the field of the pot magnet sets up in it an induced E.M.F. This E.M.F. is actually in the opposite direction to the back E.M.F. of self-induction, and so has the same effect on the impedance as if the coil possessed a certain amount of capacitive reactance. The corresponding capacity is called the "motional capacity."

The increase of inductive reactance with frequency, and the consequent increase in impedance, has the effect of reducing the current for a given voltage input to the power stage as the frequency increases. This would lead to attenuation of the higher frequencies, but it seems to be compensated for to a large extent by the fact that the conical diaphragm, in addition to its axial vibration as a whole, undergoes elastic vibrations of its material at these frequencies. The number of possible vibrations of this type is very large, and the complicated mechanical resonances produced combine to give a fairly uniform output at high frequencies

Moving coils are either wound with many turns of fine wire, in which case they are said to be "**high impedance**" coils, or with a few turns of thicker wire which can carry a larger current. The latter are known as "**low impedance**" coils. For reference purposes, an average effective impedance at about 1,000 or 800 cycles/second is usually quoted, although, as pointed out above, the actual impedance varies with the frequency This average impedance is about twice the D.C resistance of low impedance coils, and four times the D C. resistance of high impedance coils.

It was shown in Section F that, for maximum undistorted power in the output impedance, the latter should be twice the A.C. resistance of the associated valve, if a triode. It is thus of great importance to ensure that the valve and loudspeaker are correctly matched. Apart from the loss of power, higher frequencies are attenuated if the loudspeaker impedance is too large, for its varying impedance with frequency then produces a greater proportional effect in varying the total impedance of the stage. The value of transformer coupling in altering the effective impedance of the output so as to give the correct ratio, was pointed out in paragraph 47. This type of coupling also carries out the necessary function of preventing the steady anode current from flowing through the moving coil, and, if it is not employed, a choke filter circuit must be inserted in its stead

When a pentode output valve is used, the impedance of the loudspeaker will generally be considerably less than that of the valve, and so a very uniform frequency response may be expected. If the loudspeaker impedance is too low, the power output will be inefficient, and if too high, the voltage developed may damage the valve and frequency distortion may also arise. The manufacturer's instructions in any particular case should therefore be carefully followed.

55. **Moving Iron Loudspeakers.**—These may employ either a "balanced armature," "reed" or "inductor" drive.

The principle of each of these drives is the same. A small, light, soft iron armature is suspended in the magnetic field of a permanent magnet, and attached to the loudspeaker diaphragm. A coil of wire (or two coils in series) is wound round the yoke of the magnet, the ends being connected to the loudspeaker terminals.

The audio frequency (speech) currents from the amplifier pass through the loudspeaker winding, and so give rise to corresponding variations in the flux from the magnet, which, in turn, vary the force exerted by the magnetic field on the iron armature, thus causing it to vibrate together with the loudspeaker diaphragm in approximate accordance with the speech current oscillations.

In the " reed " drive, a metal reed is securely fixed at one end so that it lies close to but just not touching the poles of the loudspeaker magnet. The diaphragm is attached to the free end of the reed. Variations in the magnetic flux density cause the reed to bend to a greater or less extent, and so to vibrate about its fixed end. With this type of movement, care must be taken that if any direct current is allowed to flow through the windings it flows in the direction which increases the flux of the permanent magnet. This is usually indicated by the terminals of the loudspeaker being marked + and −.

In the " balanced armature " type of drive, the armature is suspended in the centre, and the magnets are so arranged that their initial pull on the armature is practically eliminated, the latter being affected only by the variations of the magnetic flux due to the oscillatory speech currents. This method enables the armature to be more freely suspended than in the " reed " type, resulting in more uniform reproduction and a greater response to the lower audio frequencies.

On the other hand, any direct current flowing through the loudspeaker windings will have a greater tendency to pull the armature over on to the magnet poles. Hence, with this type of loudspeaker it is always advisable to use a choke-filter or transformer coupling between the amplifier and the loudspeaker.

In the " inductor " type of loudspeaker, the iron armature is supported in a manner which allows considerably greater freedom of movement than is possible with either the reed or balanced armature type of loudspeaker. Instead of the armature moving towards and away from the poles of the magnets, it moves like a piston between two pairs of magnetic poles

The chief advantage of this type of speaker is that the freedom of suspension, combined with the large permissible movement of the armature, provides a better frequency response characteristic than that of other types of moving iron speaker. This type of loudspeaker compares favourably with an average cheap moving coil loudspeaker, both as regards sensitivity and quality of output, and has the advantages of lightness and cheapness, and also that no field energising current is required. As in the balanced armature and moving coil types, no direct current should be allowed to flow through the loudspeaker windings.

56. Loudspeaker Networks. Impedance Matching.—It is often necessary to feed a group (or groups) of loudspeakers through short lines from an amplifier. The speakers may be joined at will in series-parallel combinations, but it is necessary to take account of the effective impedance presented by the network in order to design the necessary impedance matching. Fig. 36 shows four 5-ohm speakers connected in series-parallel, the whole presenting an effective impedance of 5 ohms. This combination could be matched to the output of an amplifier by means of a transformer coupling having a suitable transformer ratio.

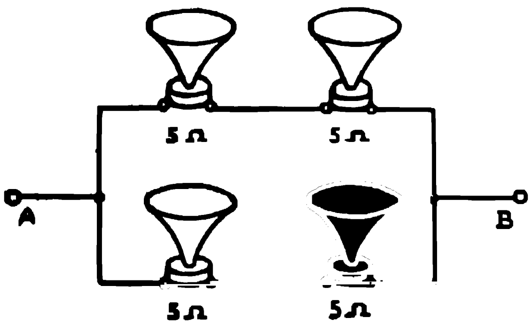


FIG. 36.

In general, there are three ruling considerations in the design of loudspeaker networks :—

(a) The impedances of all groups in parallel should be equal, unless unequal distribution of the volume of sound is definitely required.

(b) The impedance of each loudspeaker in any series leg should be equal, unless it is desired to supply some speakers with more power than others.

(c) Moreover, if all loudspeakers are to receive the same proportion of the total power, their effective impedances must be equal.

If these considerations are not observed, some speakers will have a high-pitched tone and others a low pitched one. For example, if two 5-ohms speakers are connected in series and placed in parallel

with a 10-ohm one, conditions (*a*) and (*b*) above are satisfied, but condition (*c*) is not fulfilled and the 5-ohm speakers will each receive half the power of the 10-ohm one.

It is sometimes desirable to depart from these rules in order to adjust the distribution of tone and volume among the loudspeakers. In the first instance, it is usually better to arrange speakers to give equal results, modifying the arrangement later, if necessary.

If the effective impedance of a loudspeaker is increased (or its transformer step-down ratio increased), the volume will be decreased and the tone will become more mellow (less brilliant), and *vice versa*. When adjusting the impedance of individual speakers, it is important to preserve the total effective impedance at its correct value. It may not always be possible to maintain the impedance at exactly its correct value, but, in practice, the matching will be good enough if it is within 50 per cent. of the correct value. The use of too high an impedance reduces the higher audio-frequencies and renders reproduction relatively mellow, the use of too low an impedance accentuates the higher audio-frequencies and renders the reproduction relatively brilliant. From the point of view of obtaining maximum volume, the use of too high an effective impedance has a less deleterious effect than the use of one which is too low.

Fig. 37 (*a*) represents three speakers of different impedance, each with its own transformer having a suitable step-up ratio so that the effective load presented to the load end in each case is 12,000 ohms. The total load presented to the amplifier in that case would be 4,000 ohms.

Fig. 37 (*b*) shows two speakers of different impedance, each connected to the same transformer, the latter having tappings providing the requisite transformer ratio; each speaker, singly, produces an effective impedance of 12,000 ohms across A and B, the two in parallel being equivalent to 6,000 ohms.

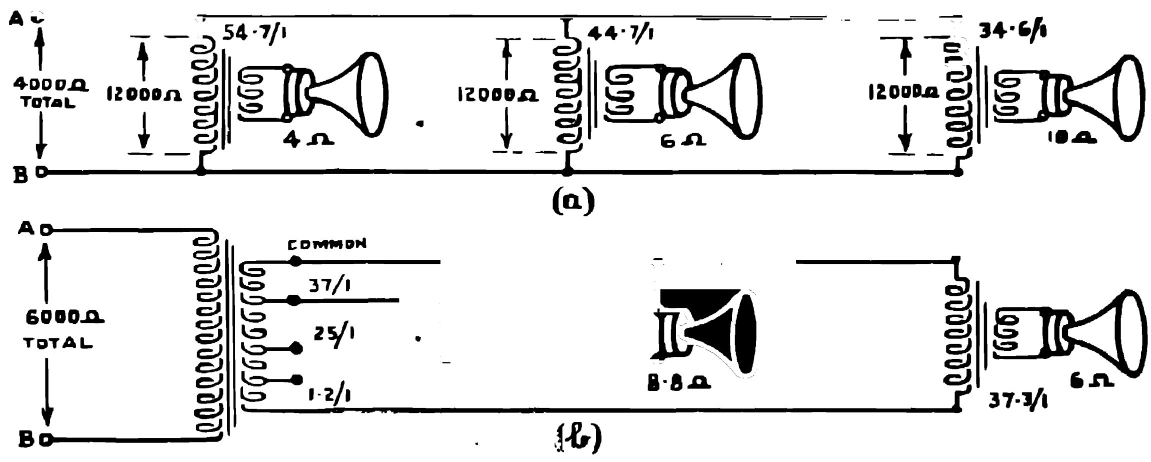


FIG. 37.

In the above discussion it has been assumed that the loudspeaker groups are in the immediate neighbourhood of the amplifier. When the distance between the amplifiers and loudspeakers exceeds 100 yards, the resistance loss in the leads is likely to be sufficient to cause appreciable attenuation of the signal, and considerable distortion may be introduced. Simple impedance matching is not sufficient, and recourse must be had to the complexities of A/F "line working."

57. **A/F Line Working.**—In practice, cable links—here called **lines**—of varying length are often required to carry currents at audible frequencies; they are required whenever it is necessary to connect up two units situated in different places. Practical instances are to be found in any "outside broadcast," or in any complex public address system, such as that organised in H.M. Dockyards during Navy Week, etc. Lines of varying length are required to connect first stage

amplifiers to main amplifiers, main amplifiers to loudspeakers, radio receivers to main amplifiers, etc. Examples of **shorter lines** include those connecting gramophone pick-ups and main amplifiers, or the photo-electric cells and first stage amplifiers of talkie installations.

In line work it is common practice to refer to the end of the line at which power is supplied as the **generator end,** and the end of the line at which power is supplied from the line to the load as the **load end.**

Lines may consist of twin core or single core cable, and, in general, the constants (resistance, inductance and capacity) are uniformly distributed along the length. For the present purposes, a *short line* may be defined as one in which the distributed inductance, resistance, and capacity, may be regarded as concentrated at one point without introducing serious errors. The attenuation produced by the line depends on the frequency, an effect which accounts for "line distortion." The effect becomes more pronounced when the length of the line becomes commensurate with the wavelength corresponding to the mean audio-frequency. It is, therefore, a matter which can be partially mitigated by taking the usual steps to prevent "line resonance," *i.e.*, matching to the surge impedance at the generator and load ends of the line respectively. The average surge impedance for twin core cable is between 150 and 200 ohms, but may be as high as 300 ohms (R.37). In general, the attenuating effect of the line may be neglected with short lines up to about 30 ft. in length, the load impedance being directly matched to the generator end. Lines longer than about 30 ft. should be made into non-resonant lines by careful matching of the impedance at each end.

58. **Short, High Impedance Lines. Effect of L, C and R.**—The three main factors affecting short line work are the series resistance, series inductance, and the shunt capacity.

The series resistance offers a uniform impedance at all frequencies; typical values for single or double core cable range from 1 to 1·5 ohms per 100 ft. The series inductance offers an impedance to the line current which varies directly with the frequency; for most samples of cable the inductance is of the order of 20 mics. per 100 ft. The shunt capacity provides a by-pass for the line current, the impedance of which varies inversely with the frequency. In the case of single core cable, an average value of capacity is 0·006 microfarad between wire and earth; it is about 0·004 microfarad between wire to wire in the case of twin core cable per 100 ft.

The magnitude of the effect of these impedances on the line currents will depend, almost entirely, on the relative size of the effective generator and load impedances.

The impedances offered by the series resistance and inductance respectively, will have less effect when either, or both, of the generator and load impedance are high in comparison. As a rough working rule, the generator or load impedance should be at least twice the impedance offered by the inductance or resistance at the highest frequency that it is desired to transmit.

The shunt capacity will have less effect when either, or both, of the generator and load impedances are low in comparison. As a similar rough working rule, one might state that the generator or load impedances should be less than half the impedance of the shunt capacity at the highest frequency that it is desired to transmit. For short lengths of cable (a few feet only), the capacity of most cable offers an impedance of not less than 500,000 ohms at 8,000 cycles/second. For lengths of cable up to 30 ft. or so, the shunt capacity is about 25,000 ohms at 8,000 cycles/second. For lengths up to about a quarter of a mile, the shunt capacity offers an impedance of about 500 ohms at 8,000 cycles/second.

> *Example* (*a*).—A line feeding a 10-ohm loudspeaker will have little effect on the reproduction provided that the resistance of the line is not more than 5 ohms, and the impedance of the inductance at 8,000 cycles is also not more than 5 ohms. What is the maximum length of cable that can be used? The cable constants (twin core) are $L = 20$ mics. per 100 ft., and $R = 2 \cdot 6$ ohms per 100 ft. of double cable.

It is evident that 200 ft. of cable has a resistance of 5·2 ohms, and an inductance of 40 mics.—giving a reactance of about 2 ohms at 8,000 cycles/second; the reactance for the double cable would amount to about 4 ohms. 200 ft. is, therefore, the greatest length of this type of cable that should be used to feed a 10-ohm loudspeaker; if a 5-ohm loudspeaker is used, only half this length is permissible. Owing to the low value of the load impedance, the effect of the shunt capacity is negligible.

Example (*b*).—A 5-ft. length of cable connects a photo-electric cell (of average impedance about 20 megohms) to the coupling resistance in a first stage amplifier. What is the maximum value of load resistance that will enable all frequencies up to 8,000 cycles/second to be transmitted without appreciable attenuation ? The cable constants are C = 0·002 microfarad per 100 ft. and L = 20 mics. per 100 ft.

The shunt capacity of 5 ft. of cable is, therefore, 0·0001 microfarad, and the reactance of this capacity at 8,000 cycles/second is 200,000 ohms. Hence, the load resistance must not exceed about 100,000 ohms, in order to prevent the impedance of the line from having a controlling effect. Owing to the high values of the load and generator impedances, and to the short length of the connecting cable, the effect of the series inductance and resistance will be quite negligible.

Fig. 38 represents a simple direct link between a generator, of effective impedance 20 MΩ, and a load of about 250,000 ohms. The line is shown as *single core* screened cable, and the reactance of the shunt capacity should not be less than about 350,000 ohms.

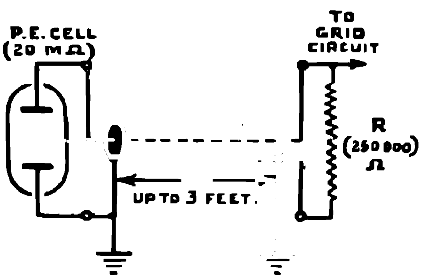


FIG. 38

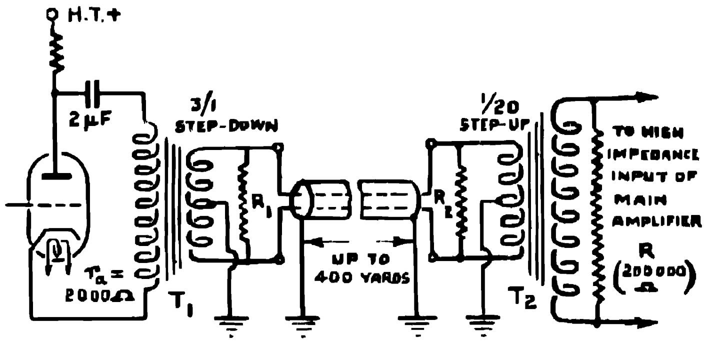


FIG. 39.

59. **Longer Lines. Low Impedance Lines.** For lines over 30 ft. in length, the impedance rapidly falls, and all such lines are best treated in the non-resonant state, a matter involving impedance matching with the surge impedance at each end. Fig. 39 represents a typical low impedance line, and gives the necessary impedance matching arrangements. The effective resistance of the primary of transformer T_2 is 500 ohms ; resistances R_1 and R_2 are 1,000 ohms each, and their effect is to lower the total shunt impedance to about 250 ohms and assist the line in maintaining a uniform frequency response

60. **Effects of Electrostatic Pick-up and Extraneous Noises.**—All lines are liable to pick-up noises from stray electrostatic and electro-magnetic fields. High impedance lines are more susceptible to this pick-up than low impedance lines. Likewise, lines in which one conductor is earthed and the other is "live" are more susceptible than twin core cables which are balanced to earth, *i.e.*, where the transformer windings at the ends of the line are centre tapped (Fig. 39), the centre tap being earthed ; or, alternatively, where the terminal impedances consist wholly, or in part, of centre tapped resistances, the centre tap being earthed.

Both electrostatic and electro-magnetic pick-up is greatly reduced by the use of screened cable, *i.e.*, single core or twin core cable surrounded by tinned copper braiding which must be well earthed at both ends (Fig. 39), and, in the case of longer lines, at several intermediate points.

It will normally be quite satisfactory in the case of high or medium impedance lines to employ single core screened cable, the metal braiding being used as an earth return as well as a screen. For the longer low impedance lines, however, twin core screened cable should be used, and the two conductors should be balanced to earth by employing transformers with earthed centre taps. The screening must, of course, also be well earthed at both ends and at several intermediate points.

61. **Long Lines.**—In the case of lines exceeding half a mile or so in length, the inductance and capacity of the cable cannot be considered as concentrated at one point but must be considered as distributed uniformly along the line, or, as an approximation, concentrated in groups at points not more than about one mile apart. The treatment of this subject is a highly specialised one, and will not be further considered here.

62. **Gramophone Pick-ups.**— A gramophone pick-up is, essentially, a device for converting the mechanical vibrations of a gramophone needle into corresponding electrical oscillatory voltages which are capable of being subsequently amplified and reproduced by loudspeakers in the form of sound. These electro-mechanical devices may be classified in a manner similar to that employed in the case of microphones and loudspeakers (paragraphs 9 and 50). Representatives exist in each of the classes, but most modern pick-ups belong to the following three types :—(*a*) moving-iron armature pick-ups, including "needle armature" pick-ups ; (*b*) moving-coil armature pick-ups ; and (*c*) electrostatic pick-ups working on the condenser microphone principle.

Fig. 40 (*a*) represents a moving-iron pick-up. A light soft iron armature (A) is pivoted at (B) in a small gap between the poles (P) of the permanent magnet. Around the yoke is a winding (C) of fine wire, the ends of which are led away to the pick-up terminals (E) and (F). A gramophone needle is secured as shown, and as it runs in the groove on the record, it produces corresponding vibrations of the armature about its pivot ; a record groove wobbles *radially* and *not vertically*, as is sometimes supposed. The vibrating armature causes variations in the magnetic flux which, in turn, give rise to corresponding back E.M.Fs. in the pick-up winding (C). In the "needle armature" type of pick-up, the needle itself performs the function of the armature.

In moving-coil pick-ups, the needle is attached to a small coil of wire pivoted between the poles of a permanent magnet, the ends of the coil being led away to the pick-up terminals. The principle is that of a moving-coil microphone (paragraph 11), the vibrations of the coil causing corresponding E.M.Fs. to be set up in the coil itself. Moving-coil pick-ups are not yet used to any great extent, and most commercial pick-ups are of the moving-iron type.

In all cases, the voltages developed across the pick-up terminals should, in the *ideal* pick-up, correspond exactly with the original vibrations cut into the record. The average pick-up has a mean output of about 0·5 to 1·5 volts R.M.S. on a normal record. Some pick-ups have an output of

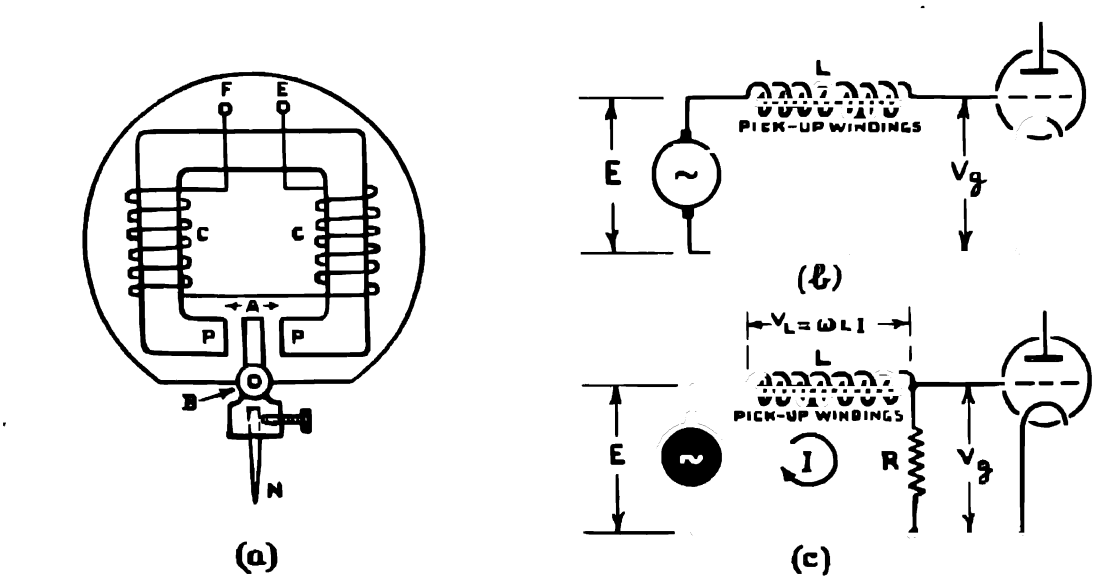


FIG. 40.

2 or 3 volts, and with these care must be taken not to overload the first valve of the amplifier; others, including needle armature pick-ups, have an output of the order of two or three hundredths of a volt only. The output from a pick-up is always applied to an amplifier, through a suitable impedance matching system; pick-ups of the latter type require more amplification than others.

TONE CONTROL WITH GRAMOPHONE PICK-UPS.—In cases where a pick-up constitutes a part of an R/T receiver, the tone controls incorporated in the latter may be made to serve also for the radiogram. A certain fixed amount of tone control is, however, usually included.

Fig. 40 (*b*) shows the pick-up connected directly to the input of a triode. It will be seen that, since the impedance between grid and filament is very large, the current flowing in the circuit will be negligible. Hence, there will be no appreciable voltage drop across the windings, represented by inductance L, and the voltage V_g applied to the grid of the valve will be the same as the voltage E produced in the pick-up; this will be approximately true at all audio-frequencies.

Fig. 40 (*c*) shows the same circuit with a shunt resistance R connected across the pick-up terminals. A current will now flow round the circuit, and there will be a voltage drop across the inductance L. Under these conditions, the voltage V_g will be less than E. Moreover, since the voltage drop across the inductance L will be greater for the higher audio-frequencies than for the lower ones, it follows that, for a given value of E, the value of V_g will be smaller at the higher audio-frequencies than at the lower ones. Hence, the upper musical register will be reduced to a greater extent than the lower one, and the tone will become mellow and less harsh. If R is too small, the upper musical register will be reduced so much that definition will be lost, brilliance will be lacking, music will be woolly and speech drummy and indistinct, the consonants and sibilants being almost entirely absent (paragraph 42).

The extent to which the higher audio-frequencies are reduced will also depend on the value of the inductance L. Other things being equal, the greater this inductance, the greater will be the attenuation. That is to say, a given value of shunt resistance R will cause a greater reduction of the upper musical register with a high inductance pick-up than will be the case with a pick-up of medium or low inductance. Pick-ups having an inductance of about 5 henries are usually known as "high impedance" pick-ups, those with windings of inductance about 0·5 henry being known as "medium impedance" pick-ups.

A similar but more marked effect is produced if a small capacity is shunted across the pick-up terminals in place of the resistance R, but in this case a resonant peak is formed at a frequency determined by this capacity and the inductance of the windings. The resonant frequency is lower with high impedance pick-ups than with medium impedance ones; this resonance may sometimes be used to produce a pick-up having uniform response up to the resonant frequency. Shunt capacity may be introduced intentionally, as in some needle scratch filters, or it may be introduced unintentionally by the use of long leads between the pick-up and the amplifier, or by the use of a pick-up transformer of too high a ratio.

From the above it is obvious that a variable resistance connected across the terminals of a moving-iron pick-up may be used as a form of variable tone control.

NEEDLE SCRATCH.—Needle scratch is caused by the needle point rubbing over the fine abrasive compound contained in the surface of the record. Its particles impart a series of shocks to the needle at small and irregular intervals. Analysis of the "noise" shows that it consists mainly of frequencies between 4,000 and 6,000 cycles per second, extending downwards to 3,000 or 2,000 cycles per second with old records. The frequencies in the neighbourhood of 3,000 to 4,000 cycles are responsible for a coarse, harsh scratch, those between 4,000 and 6,000 cycles producing a finer hissing noise.

Needle scratch is greatly accentuated by the armature resonance which exists in most pick-ups; the needle and armature have a natural frequency of oscillation, often at about 4,000 cycles per second. The nett result is a marked increase in the strength of harsh needle scratch, with predominating frequencies in the neighbourhood of 4,000 cycles per second.

High note attenuation affords a partial cure for needle scratch, but electrical filter circuits are also used, the object of which is to produce a *dip* in the frequency response characteristic at the same frequency as the *peak* caused by armature resonance. If the dip is equal in magnitude to the peak.

the overall result will then be a tolerably uniform response characteristic. A dip in the characteristic can be produced by connecting an acceptor circuit, tuned to the required frequency, across the pick-up terminals.

63. **Typical Modern Superhet Broadcast Receiver.**—Fig. 41 represents the simplified theoretical diagram of an R/T receiver including most of the features typical of the cheaper modern broadcast sets. The receiver is designed for all-mains operation and is provided with five valves, including rectifier. The aerial is inductively coupled to a band-pass filter having two tuning ranges; the filter employs a mixed coupling in order to keep the band of frequencies passed constant in width at all frequencies. The band-pass circuit feeds a heptode frequency changer (1), which is controlled by the A.G.C line, the output being coupled through a 125 kc/s. I/F band-pass circuit to a variable mu screen grid I/F amplifying valve (2),also controlled by the A.G.C. line. The output from valve (2) is I/F band-pass coupled to a double-diode-triode (3), which combines the functions of second detector, first A/F stage, and also provides the A.G.C. negative bias. This stage is resistance-capacity coupled to the high slope pentode output valve (4), designed to give an output of about 2 watts in the case of most domestic R/T receivers. In this case, provision is only made for amplified delayed A.G.C., but Q.A.G.C. could easily be provided, in addition, in the way indicated in Fig. 30.

The receiver could be designed with four controls—(*a*) a single knob tuning control; (*b*) a volume control, operating both on radio and gramophone; (*c*) a master switch—gram., M.W., L.W., and off; and (*d*) a tone control operating both on radio and gramophone. The two frequency ranges which are normally provided are the 150-300 kc/s. range (2,000-1,000 metres, and usually labelled L.W.), and the 545-1,500 kc/s. range (550-200 metres, usually labelled M.W.). In Service nomenclature this covers most of the M/F wave band, with a gap in the middle. Many receivers now contain a third range to include the stations working on H/F.

The aerial input is mutually coupled to the filter by coils (5) and (6), provision being made for the necessary increase in coupling at the lower frequencies. Condenser (7) may have a value of about 0·0005μF, and its object is to reduce the effective capacity of the aerial, thereby rendering the condenser (8) more effective as a tuning control. With the ordinary commercial receiver, it is most desirable that the tuning should be as independent as possible of the nature of the aerial to which the set is attached. If the capacity of condenser (7) is increased, the strength of the signals will also increase; at the same time the aerial load will increase, and the nett effect will be to make the whole circuit less selective. The smaller the condenser the more selective will be the circuit; the lowest practical size of this condenser is about 0·0001μF.

Coils (9), (10), (11) and (12), with large condenser (13) of capacity 0·06μF, constitute the **mixed coupling** of the band-pass filter. The mutual inductive coupling is most operative on the higher frequencies, the *common capacity* coupling (13) is most operative on the lower frequencies, and the nett effect is a partial levelling out of the coupling over the whole frequency band; with suitable care and design it is possible to pass only a band of frequencies of constant width. Additional coupling at the low frequency end of the range is provided by coil (14). For the wave ranges indicated, the maximum value of condensers (8) and (16) (and the similar condenser in the secondary circuit) will usually be 0·0005μF, with a coil of about 180 mics. for one range, and one of about 1,860 mics. for the other. The condenser must produce an effective variation in capacity of 10 : 1 on the M.W. band, and 4 : 1 on the L.W. band.

The heptode frequency changer (a tetrode in series with a triode oscillator) employs the normal circuits, the electron coupling being used to "mix" the two frequencies. The self-oscillator circuit (15) and (16) consists of an ordinary "reversed feed" oscillator, with a tuned circuit between grid and cathode. Condenser (16) is the third condenser of the gang; for medium wave frequencies, the lower half of coil (15) is short circuited. Components (17) and (18) constitute a grid leak and condenser combination which improves the efficiency of the oscillator by providing a small negative bias on the oscillator grid, and also helps to stabilise the frequency (R. 38). The anode feed resistance (19) provides the requisite P.D. between the oscillator anode and earth. Resistance (20) in series

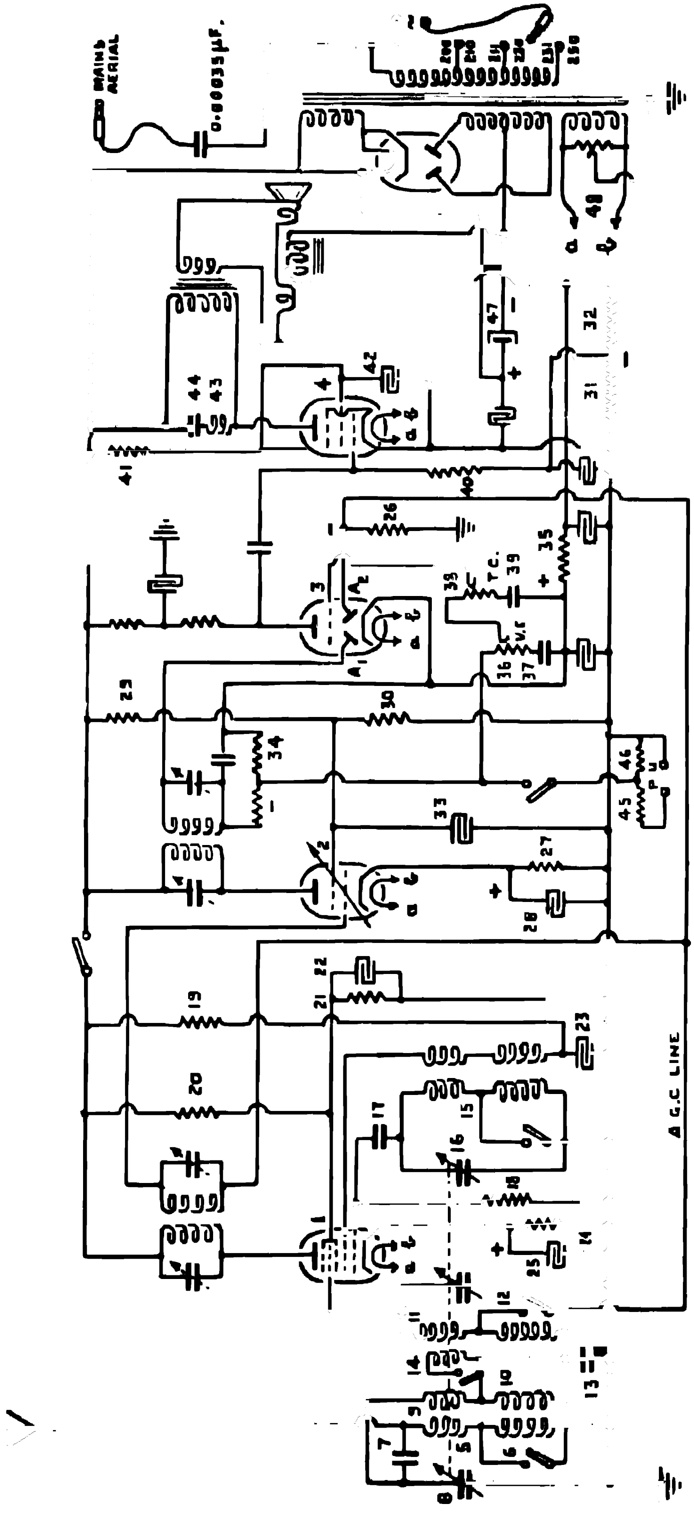


FIG. 41

with (21), (31), and (32), forms a potentiometer across the H.T. supply, and maintains the screen of the tetrode at the requisite potential; resistance (21) is by-passed by condenser (22) to prevent any high frequency P.D. from existing between the screen and cathode. Components (19) and (23) together form a decoupling resistance and condenser, and components (24) and (25) supply the heptode with automatic negative grid bias, by maintaining the cathode slightly positive with respect to earth.

The heterodyne frequency produced by the triode oscillator is always higher than the signal frequency, and the frequency difference is adjusted to be that to which the I/F amplifier is tuned. For example, when receiving London Regional on 877 kc/s., the triode oscillator would be set at 902 kc/s., giving an intermediate frequency of 125 kc/s. It may here be noted that an **image signal** from an unwanted station on 1,027 kc/s. can also produce the same intermediate frequency and would therefore be received, unless very highly selective circuits are used. In certain cases, special arrangements are made to suppress **second channel interference** of this nature.

The tuned transformer band-pass output impedance applies the band of intermediate frequencies between grid and cathode of the variable mu screen grid valve. The feed to the control grid is direct, the feed to the cathode being through the A.G.C. bias resistance (26). Automatic grid bias at this stage is provided by components (27) and (28), the control grid being also joined to the A.G.C. line. Resistances (29), (30), (31) and (32) together constitute a potentiometer across the H.T. supply, and serve to supply the necessary positive potential to the screen grid; condenser (33) acts as R/F by-pass.

The output from valve (2) is applied to the three-purpose valve (3) by the tuned transformer band-pass output impedance. The cathode and anode A_1 constitute a diode, and their attached circuits are of the conventional type described more fully in paragraph 27. The A/F output is obtainable across the load resistance (34), and is applied between cathode and grid of the triode portion of the valve. The lead to the cathode is direct, but that to the grid lies through the two parallel circuits constituting the **volume control** and **tone control** respectively. The amplified A/F signals are passed on to the pentode output stage through a resistance-capacity coupling.

Delayed amplified A.G.C. is obtainable by means of the circuits attached to the cathode and anode A_2. These are the same as those described in Fig. 30. With no input signals, the returning space current flowing through resistance (35) makes the cathode positive with respect to the anode A_2 which is earthed. Resistances (31) and (32) form part of a potentiometer across the H.T. supply and correspond with resistance (2) of Fig. 30. No A.G.C. grid bias is available across (26) until the signal input exceeds the critical value which is determined by the magnitude of the delay voltage.

The volume and tone controls are circuits of essentially the same nature connected in parallel across the load resistance (34). There is, however, the difference that the P.D. available across the VC circuit is arranged to be substantially independent of the frequency for all settings of the control, while that available across the TC circuit varies with the audio-frequency.

The volume control consists of a variable resistance (36) of value 0·5 megohm in series with a fixed condenser (37) of value 0·1 microfarad. With full resistance, there is sensibly no difference between the impedance of the circuit at 1,000 and at 10,000 cycles per second; with no resistance, the reactance of the condenser only varies from 160 ohms at 10,000 cycles t[illegible] ohms at 1,000 cycles. The tone control circuit consists of a variable resistance (38) 0·5 n[illegible] cap[illegible]n series with a fixed condenser (39) 0·0005 microfarad. In this case, with full resistance th[illegible] is an appreciable difference between the impedance of the combination at 10,000 cycles and at 1,000 cycles, the values being roughly 0·5 megohm and 0·6 megohm respectively. With no resistance the difference is relatively very considerable, being about 0·03 megohm at 10,000 cycles and 0·3 megohm at 1,000 cycles; this setting of the tone control would be attended by drastic attenuation of the higher audio-frequencies, for any setting of the volume control.

The input to the high slope pentode output valve is quite normal, resistance (40) constituting a grid leak, the grid being held at a small negative potential with reference to the cathode, which is

earthed by means of the potentiometer resistance (31). Anode feed resistance (41) supplies the necessary potential to the priming grid, the latter being held at the same oscillatory potential as the cathode by means of condenser (42). The output is transformer coupled to a low impedance moving-coil loudspeaker, the field coils of the latter acting as a smoothing choke in the rectified D.C. supply. Across the primary of the step-down transformer is a tone control circuit (43) and (44), the function of which is to attenuate the higher audio-frequencies which tend to be over accentuated with the high impedance pentode output.

Provision is made for the **gramophone pick-up,** and resistances (45) and (46) constitute a fixed tone control (paragraph 62).

The rectifier is an ordinary full wave circuit, condenser (47) being of the large electrolytic type and of value $8{\cdot}0\mu F$ Heating current to the filament is obtained from the winding *ab*, a **hum dinger** (48) also being provided. The latter consists of a potentiometer, of value about 50 ohms, connected across the heater supply, the sliding contact enabling the earth connection to be applied at any point. In any model, owing to constructional difficulties, the lead from the grid of an A/F stage is likely to be nearer to one heater lead (say *a*) than to the other. By capacitive coupling the nearer heater lead will tend to affect the potential of the grid more than the further one, and the object of the adjustment of the hum dinger is to increase the effect of the further lead, which is in anti-phase, until the two effects balance out. In practice, it is seldom possible to eliminate all hum, but it can be considerably reduced.

For the purposes of simplification, all screening trimming condensers and details of switches have been omitted from the diagram. Screens should surround the tuned transformer I/F output stages, and screened cable should be used for the aerial input, the lead from the band-pass filter to the heptode, and in the A/F input lead to the triode portion of valve (3) ; the leads from the pick-up should also be screened.

64. **Sensitivity, Selectivity, Fidelity.**—The relative behaviour of different **broadcast receivers** may be assessed numerically and expressed in terms of their sensitivity, selectivity, and fidelity.

SENSITIVITY.—This represents the relative ability of receivers to respond to small input signal voltages. It has been arbitrarily defined as the signal input, usually expressed in microvolts, which must be developed in a standard aerial in order to produce an output of 50 milliwatts, when the carrier is modulated 30 per cent. at a frequency of 400 cycles.

Fifty milliwatts is a very small output for loudspeaker purposes, and was chosen originally because it is considered to be the smallest one to have any loudspeaker programme value (App. A5). A set of sensitivity 100 microvolts would not give satisfactory reception if that were also the value of the signal voltage input ; an output of about 1 watt is to be preferred (twenty times the conventional 50 milliwatts), for which an input of $100 \times \sqrt{20} = 447\mu V$ will be needed.

Sensitivity is measured with the volume control set for maximum volume, and the receiver tuned to give maximum response at the frequency in use. The signal is injected by means of a calibrated signal generator, now usually part of the equipment of any radio service engineer. In order to have a basis of comparison which may represent the performance of the receiver when attached to an aerial, the signal generator is not attached directly to the set but applied in series with a standard **dummy aerial,** having the constants $L = 20\mu h$, $C = 200\mu\mu f$, and $R = 25$ ohms.

The sensitivity of a receiver cannot be usefully increased beyond a certain limit unless the selectivity also becomes greater. If the sensitivity is strong enough to give reception of weak stations previously missed, the selectivity must be great enough to keep out the powerful local stations. It is in this respect that the superheterodyne type of receiver scores over the older "straight" sets. The presence of much **man-made static** provides a second reason setting a limiting value to the useful sensitivity of a receiver. In big cities, the maximum usable sensitivity is of the order of $100\mu V$, if the requisite signal/noise ratio is to be maintained (App. A9).

Some idea of the sensitivity of various receivers may be gained from the following table of average values :—

Type of Set.	Maximum Sensitivity.
Simple sets, consisting of a detector and output valve	100,000 to 200,000 μV.
Straight sets, with one R/F stage	2,000 μV.
Straight sets, with two R/F stages	20-100 μV.
Simple superheterodyne sets	50-250 μV.
Superheterodyne sets, with one A/F stage	5-20 μV.
Superheterodynes for use in cars, etc.	About 1 μV.

Most of the cheaper superhet broadcast receivers have sensitivities between 50 and 100μV. Since efficient amplification becomes more difficult as the frequency increases, the sensitivity of a set usually decreases with the frequency and particular receivers may often have bad blind spots covering a band of frequencies at which the sensitivity is very low.

Care must be taken not to make unfair comparisons between the sensitivity of R/T receivers designed for loudspeaker reproduction and W/T receivers built for reception on telephones. The **sensitivity basis of comparison for W/T receivers** is usually quite different ; the sensitivity of Service receivers is defined by the input required to give one milliwatt output after two stages of note magnification, using a carrier modulated 80 per cent. at 400 cycles/second. Under those conditions it is found, in general, that the receiver gives a good R9 signal at the detector. In some tests with a Service W/T receiver, covering frequencies from 140 to 1,500 kc/s., the sensitivity figures range from 30μV at the best to 12,000μV at the blind spots on the highest range. Using the **broadcast standard sensitivity basis** of reference, the above Service receiver, at its best, gives a sensitivity of $30 \times \frac{3}{8} \times 50 = 80 \cdot 0\mu V$; the comparison is, however, an unfair one, since the receiver does not include the power A/F amplifying stages, and, on phones only, the Service receiver would show a figure relatively very high.

SELECTIVITY.—The *overall* selectivity of a receiver is the mean of that due to its various circuits, and is a measure of its capacity to discriminate between wanted and unwanted frequencies. In the laboratory it is measured, at particular typical frequencies, by varying the input carrier frequency on either side of that to which the set is tuned, and adjusting the magnitude of the input signal to give the same output in all cases.

Full results can only be expressed in the form of curves, but small characteristic pieces of information may sometimes be selected from the curves and quoted instead ; thus, if the requisite inputs at resonance and 9 kc/s. off it are 10μV and 100μV respectively, the requisite input may be said to be **10 times up** (20 db.) at 9 kc/s. Selectivity may thus be given as a ratio of sensitivities at a stated frequency from resonance.

When a superheterodyne receiver employs a low I/F, the selectivity would be expected to be at its worst at the **image frequencies**, and information referring to those frequencies (the **second channel ratio**), would therefore be most valuable. A Service receiver employing an I/F of 24 kc/s. was tested in this way. In that case, the image frequency was 48 kc/s. *more* or *less* than the signal frequency, depending upon whether the first heterodyne is tuned to 24 kc/s. *above* or *below* the signal. It was found, as would be expected, that there was a progressive decrease in selectivity towards the higher frequencies, since the constant frequency difference of 48 kc/s. becomes a smaller fraction of the signal frequency On the most favourable frequency (250 kc/s.), an interfering signal on the image frequency would have to have been 60 dbs. (1,000 times) stronger than the wanted signal to give the same output. At the worst point (about 1,500 kc/s.) the difference was only 20 db. (10 times). Tests were also made on the same receiver at frequencies close in to the wanted frequency they showed that for signals about 1 kc/s. off the tuning point, the necessary

increase of strength was 6 dbs. (roughly twice as strong), while signals at about 5 kc/s. away from resonance had to be 40 dbs. (100 times) stronger than the wanted signal before giving the same output.

In the above Service W/T receiver tests, an increase of 6 db. corresponded roughly to one signal strength on the "R" scale. Including **signals one signal strength down** on that of the wanted frequency, the **band width** passed by the receiver amounted to 2·2 kc/s. Including signals 40 db. down, the band width passed was 9·2 kc/s. Selectivity is sometimes defined as "the slope of the response curve, expressed in decibels/kilocycles per second, of the portion of the curve outside the band width of the receiver." In the case of the flat topped band-pass filters used in R/T receivers, the slope of the curve often amounts to a value of the order of 60 db. per kc/s.

It may be noted that "**condenser tuning**" usually produces circuits the selectivity of which decreases as the frequency increases For R/T it is advantageous that the selectivity should be the same at a fixed number of kc/s from resonance, whatever the resonant frequency may be; with "**permeability tuning**" (F 20), the selectivity of the receiver increases with increasing frequency, giving a result which is an approach to the ideal for R/T work.

FIDELITY.—This refers to the accuracy with which the various A/F modulating frequencies are reproduced by the receiver; fidelity curves (App A8) are a measure of the frequency distortion Fidelity usually varies with the carrier frequency, and is measured in the laboratory by observing the change in output as the modulating frequency is altered, the percentage modulation being constant. The perfect fidelity curve would be a straight line parallel to the X-axis.

65. **Privacy Systems of R/T.**—Very many methods have been devised to render difficult the illicit receipt of R/T messages. Simple methods, such as wobbling the carrier wave, or inversion of the sideband spectrum, do not provide a complete solution of the problem. A method known as "scrambling"—based on frequency transposition by heterodyne action—is much more satisfactory, but the equipment involved is highly complex. It is not proposed to discuss any of these privacy methods in detail.

66. **A Modern Low Powered H/F Broadcast R/T Transmitter.** Fig. 42 represents, diagrammatically, the electrical details of an R/T transmitter of low power (about 100 watts in the aerial), designed to use an R/F carrier-frequency of the order of 3,500 kc/s. The carrier is provided by a master circuit of conventional series-fed Hartley type; using a low impedance valve having an anode dissipation of 25 watts, condenser (1) had the value of 1 jar (max), condenser (2) is 120μμF, and grid leak (3) is 15,000 ohms. The output from the master is loosely capacity-coupled through condensers (4), and condenser (5), to the control grids of two pentodes (6) and (7), arranged in push-pull. The pentodes have an anode dissipation of 75 watts, and the push-pull combination constitutes the main amplifier.

Condenser (5) has a maximum value of 100μμF, and determines the voltage output from the master which is applied as input to the main amplifier. If this condenser is made too big, its reactance becomes relatively small; the oscillatory voltage across it will be much reduced and the input volts to the amplifier will also be small. Coupling condensers (4) have values of about 45μμF; in general, the figure is governed by the type of valve used as master. (*Cf.* K.28.)

The arrangements for providing screen grid bias and the H.T. positive potentials on the anodes of the pentodes are quite standard in form, and need no explanation.

Suppressor grid modulation (N.24) is employed. In the diagram, the A/F input is shown applied through a 200 ohm line to the primary of a 5 : 1 step-up matching transformer. The transformer ratio is determined by impedance matching considerations, and by the requisite peak value of the A/F modulating voltage. In this case, a peak output from the transformer of about 225 volts was required for modulation, and it will be assumed that the line is delivering an A/F input of 45 volts peak value, representing an output of about 5 watts, to the 200 Ω line, from the microphone amplifier; in these circumstances a 5 : 1 transformer will be required. In addition to providing the requisite voltage step-up, the transformer is required to match an input impedance of 5,000 ohms

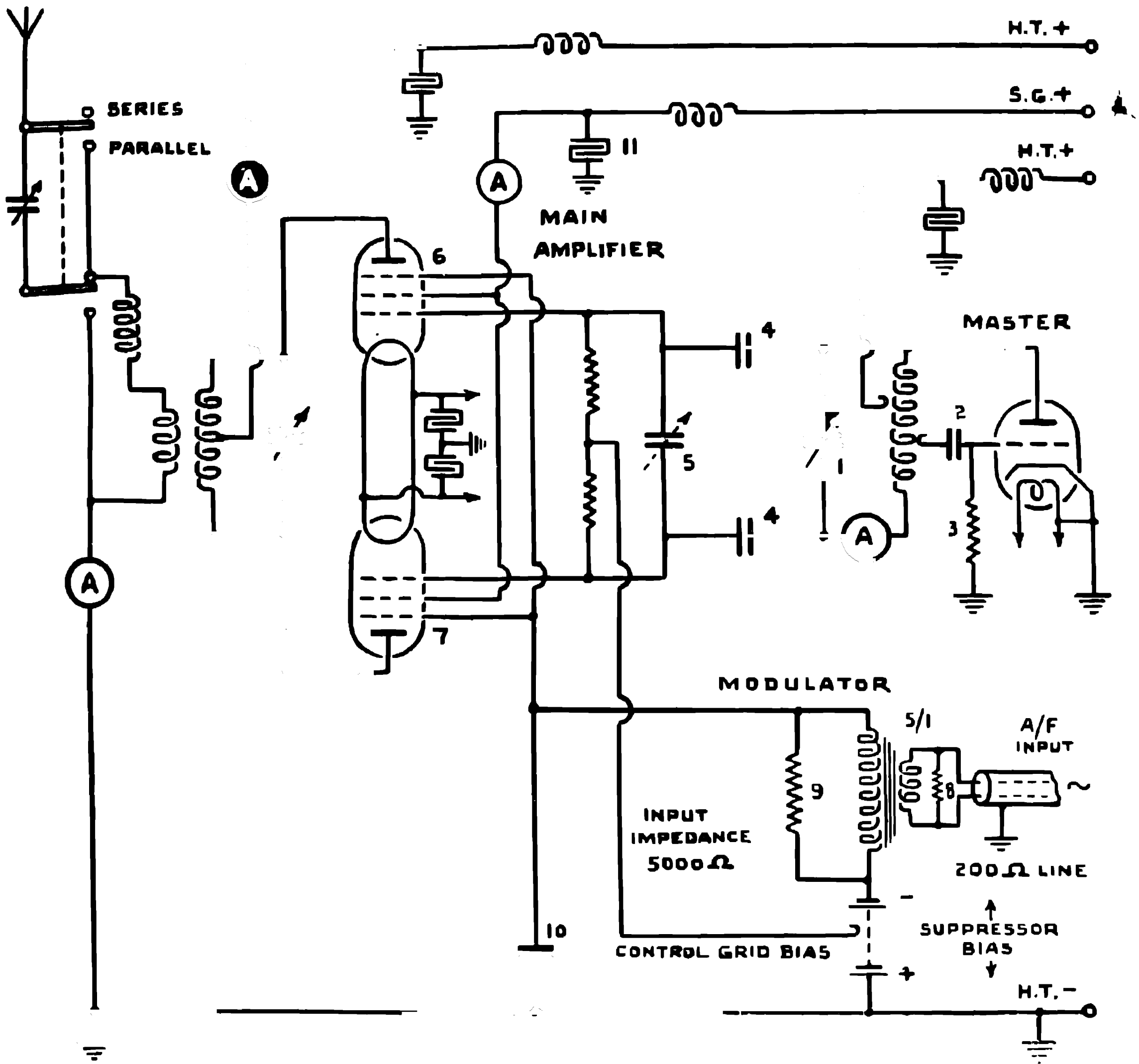


FIG. 42.

to a line impedance of 200 ohms; resistances (8) and (9) together lower the shunt impedance of the line to about 200 ohms, and enable impedance matching to be achieved using a transformer, the step-up ratio of which has been partially fixed by the considerations outlined above. In this practical case the resistance (8) had the value 600 ohms, and resistance (9) was 7,500 ohms. The A/F modulating voltages are applied to the suppressor grids in series with a steady negative bias potential. The value of the latter will depend upon the characteristics of the pentode in use; in this practical case it amounted to 250 volts negative. It will be observed that the suppressor grid bias battery is also used to provide the requisite bias for the control grids. As already explained in paragraph 24, condenser (10) must not be too large, if attenuation of the higher modulating frequencies is to be avoided; in this case a value of about 2·5 jars is suitable. This may be compared with the value of $0 \cdot 1 \mu F$ for condenser (11).

The output from the push-pull combination is applied by a magnetic coupling to the aerial circuit. The latter is provided with a condenser which may be joined in series or in parallel, optimum tuning being indicated by maximum current in the aerial ammeter; the object of this arrangement is to make the whole aerial system electrically equivalent in length to an odd number of quarter wavelengths (R.20)

SECTION "N."

EXAMINATION QUESTIONS.

1. How may speech-modulated H/F currents be produced in an aerial? Give a circuit diagram and explain its action. If a pure tone of frequency 500 cycles is impressed on the microphone, and the carrier frequency is 100 kc/s., show that the modulated current contains frequencies of 99,500 and 100,500 cycles/second.
(C. & G., Final, 1936.)

2. Explain the terms "modulated wave," "carrier wave" and "side bands" in use in wireless telephony. Give a diagram showing the apparatus necessary to convert a speech modulated wave into sound.
(C. & G., I., 1927.)

3. What is meant by a "modulated wave"? Describe the "anode choke" method of modulating the high frequency currents in an aerial by sound waves.
(C. & G., Final, 1927.)

4. Explain clearly what you understand by the expression "side band frequencies of a modulated wave." With the aid of diagrams, describe the action of an "anode modulated, Class 'C' Amplifier."
(W/T. I, (Q), 1937.)

5. Why is modulation of a carrier wave necessary for transmission of speech frequencies? Describe how modulation is achieved by one of the following methods: (*a*) Choke, (*b*) Series or Class "C." What advantages has (*b*) over (*a*)?
(Warrant Tel., (Q), 1936.)

6. Explain the action, with the aid of a circuit diagram, of either :—
(*a*) the Heising constant current method of modulating, or
(*b*) the anode modulated Class "C" amplifier.
State the reasons which make the employment of (*b*) preferable to (*a*).

7. Why are Service W/T receivers generally unsuitable for high quality broadcast reception?
(Warrant Tel., (Q), 1935.)

8. Explain with diagrams, two methods whereby a valve transmitter can be modulated by speech currents. What is meant by depth of modulation, or percentage modulation? How can this be measured in an actual radio-telephony transmitter?
(C. & G., Final, 1930.)

9. Describe with diagrams, how a single side band radio telephony transmission is produced.
State how the improvements in transmission due to single side band working are distributed :—
(*a*) At the transmitter.
(*b*) At the receiver.
What is the theoretical improvement in each case in decibels?
What difficulties arise in applying single side band working to short wave services, and how can they be overcome?
(C. & G., Final, 1937.)

10.—(*a*) Describe the action of the carbon and condenser microphones, and sketch a circuit by which either may be connected to a modulator valve.
(*b*) Explain the meaning of side bands, and draw practical conclusions concerning the design of R/T receivers.
(Qualifying for Lieut. (S), 1931.)

11. **Explain the action of a radio receiver when used to convert radio-telephony signals into audible speech signals. How does the sharpness of tuning of the receiver affect the quality of the speech output?**
(C. & G., I., 1932.)

SECTION "N."

12. Discuss the principal causes of the imperfections in the reproduction of music which are likely to occur in a wireless broadcast receiver, but omit the acoustic imperfections of the loudspeaker.

(I.E.E., Nov., 1929.)

13. Describe, with a sketch, the construction of a moving-coil type loudspeaker. Indicate in a diagram the direction of the magnetic flux in the magnetic system. On what does the force applied to the cone depend?

(C. & G., Prelim , 1935.)

14. Sketch, roughly to scale, any form of telephone receiver or loudspeaker. Give some idea of the magnitude of the currents and fluxes in operation. In what way does the motion of the moving part affect the impedance of the instrument?

(L.U., 1931.)

15. Describe, with sketches, the construction of a high-power cooled anode valve for a short wave transmitter.

Give a circuit diagram with a brief description of the output stage of a radio telephone transmitter designed to work on short waves. The output of the transmitter is 20 kW. carrier power, which is fed to a balanced transmission line of 500 ohms impedance. If the transmitter is modulated 100 per cent., what is the maximum peak voltage across the output?

(C. & G., I., 1937.)

16. A carrier frequency voltage, $e = A \sin \omega t$, is modulated by a low frequency voltage, $e = B \sin pt$ The depth of modulation is M. If the modulator has a square law characteristic, derive an expression for the output from the modulator in terms of carrier and side bands.

(C & G , I., 1937.)

PROPAGATION OF ELECTROMAGNETIC WAVES.

1. **The General Medium of Propagation.**—In Section " R " it is shown that H/F currents in aerials produce a detachment of electromagnetic energy in the form of waves of the same nature as light, a process usually called " radiation," and one which may be directional in nature.

This section is mainly concerned with a brief discussion of the phenomena that occur in the media between transmitter and receiver. The account is in general terms, and it should be realised that not all of the processes are well understood, and a more precise statement of many of them would involve concepts beyond the range of this book.

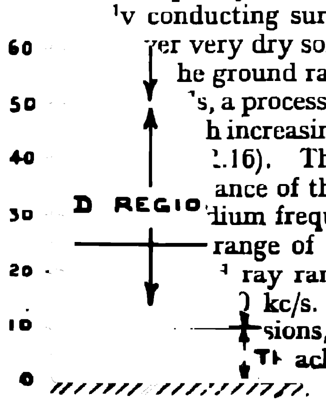
The difficulty of conceiving a wave motion without a medium in which it is passed on from point to point, had led to the postulation of a non-material medium called the æther, which must be assumed to permeate all space. It should be emphasised that properties similar to those of material substances cannot be attributed to the æther, and the attempt to do so produces absurd results. The conception of the æther is merely another way of saying that space, empty of all material substances, still possesses the property of allowing electromagnetic waves to travel through it.

When it is considered as the vehicle of transmission of electromagnetic vibrations, empty space is generally known as " free æther." All electromagnetic waves, no matter what may be their frequency, travel at the same speed in free æther. This velocity, as determined experimentally, is very nearly 3×10^8 metres per second, this figure being generally employed in wireless calculations. The more accurate figure is $2{\cdot}9982 \times 10^8$ metres per second.

When electromagnetic waves are travelling through a material medium, *e.g.*, a brick wall it must still be considered that the actual medium in which they are propagated is the æther. It is, however, no longer free æther, but is modified by the presence of the material occupying the same space. As a result, the transmission of the electromagnetic waves is also modified by the material, as in the case of light waves passing from one medium into another.

2. **Ground and Sky Waves.**—In general, an aerial does not radiate energy at the same rate in all directions of three-dimensional space ; for example, in R.40 it is shown that an earthed $3\lambda/4$ aerial, produces the most intense field at an angle of 45° to the horizontal. For the purposes of analysis, the total energy radiated into space may be divided into a " ground ray " and a " sky ray " ; these are sometimes known as the " direct ray " and the " indirect ray " respectively. The direct ray is the one which follows the surface of the earth ; the indirect ray represents the remainder of the energy, which travels at varying angles upwards prior to its possible reflection downwards from some reflecting medium. Important use is made of the latter ray in H/F transmission.

As the distance from the transmitter increases, the ground ray—if present—grows weaker and weaker, the actual strength depending very much on the nature of the intervening surface of the eart[illegible] the frequency in use. For example, a station having a ground range of 1,000 miles [illegible]y conducting surface, would have a range of 900 miles over sea-water, and one of [illegible]er very dry soil. Moreover, broadly speaking, the lower the frequency employed, [illegible]he ground ray range. This fact has been explained in terms of the tilting of the [illegible]s, a process which tends to make a wave follow the surface of the earth, provided [illegible]h increasing distance from the transmitter, is commensurate with the curvature [illegible].16). The tilting is greater the higher the frequency, a fact which accounts [illegible]ance of the ground ray at H/F.



[illegible]dium frequencies, an increase in the power of the transmitter is accompanied [illegible] range of the ground ray ; this effect is not always observed on H/F. For [illegible] ray range of Rugby at 16 kc/s. is some thousands of miles ; whereas, a [illegible] kc/s. will only have a ground ray range of a few miles. The B.B.C. [illegible]sions, employing arrays giving low angle radiation, and using frequencies [illegible] achieve communication at great distances, but cannot, normally, be

[illegible] H/F transmission on low power, in spite of the disappearance of

the ground ray, that led to the discovery of the indirect ray and the various reflecting layers in the portion of space which has since become known as the "ionosphere" (Section R.28).

3. **Various Media—The Ionosphere.**—Modern research has shown that, considering various levels with reference to the earth's surface, certain features characterise particular average heights above it, and make it clear that, in various ways, space is stratified or divided into media with distinct properties.

Up to an average height of about 10 miles, the actual height being greatest at the Equator and least at the Poles, the composition of the atmosphere remains approximately the same as at the surface of the earth, since air currents (winds) continue up to that height. The whole belt is known as the TROPOSPHERE, or region of change. The density of the atmosphere naturally falls off as the height increases, and the temperature falls to a minimum of about—68° Fahr. Above a certain critical height known as the "tropopause," the temperature remains uniform throughout a narrow belt, and later begins to increase. The tropopause marks the top of the troposphere and the beginning of the STRATOSPHERE, or region of calm. In the stratosphere, no intermixing takes place due to air currents, and the composition of the atmosphere varies with height; it becomes stratified, the lighter gases being present in greater proportions as the height increases. There is very little definite knowledge of the composition above a height of 50 miles. It probably contains a large percentage of helium. Above the top of the ozone layer, which is estimated to be at about 25 miles, definite inversion of the temperature gradient takes place and the warm belt begins; at about 40 miles the temperature is estimated to be about 70° Fahr. Prof. Piccard's excursions into the stratosphere have, so far, been confined to its lowest strata.

In considering wireless wave propagation, the electrical properties of the atmosphere are of more direct interest than its composition. Electrically, the belt from 12 to 50 miles above the earth is called the "D" region, and in it there appears to be at least one semi-conducting layer at a variable height. Generally speaking, gases conduct when they are "ionised," which means that their neutral molecules have been broken up into atoms with opposite electrical charges, possibly with the production of free electrons Ionisation involves the action of some ionising agent, and it has been thought that in the lower atmosphere it may mainly be due to the action of cosmic rays, which appear to be penetrating waves produced in the depths of space and reaching the earth from all directions; they are waves of the highest frequency so far discovered.

Against this view must be set the fact that the height of the "D" region layers is subject to a diurnal variation; one would not expect this result if the prime cause is cosmic ray action. It has been considered, however, that, if the latter cause is supported by something similar to thunderstorm action, there would be produced a sufficiently dense ionic layer to act as a reflector for waves of low frequency. Recent observations of the received echoes of short duration wireless impulses leave little room for doubt of the existence of at least one ionic reflecting layer within this regio

In 1902, two scientists, Kennelly in the United States, and Heaviside in England, ind
predicted that high above the earth there must be a densely ionised layer which acte
for wireless waves and turned them earthwards again. Since then their conjecture
justified, and a densely ionised layer has been detected at a variable height bet
90 miles above the earth's surface; this is called the "E" region, and the
KENNELLY-HEAVISIDE layer. The word "layer," however, is rather
has no sharp boundaries. The ionisation in it is mainly due to the presence
cause is still obscure, but the main factor appears to be the ultra-violet r
Such radiation is most potent at the outer edge of the atmosphere, but the
electrons it produces also depends, of course, on the number of molecules w
ionised. This number increases as the height above the earth decreases
radiation is also being absorbed as it penetrates the atmosphere. The res
density of free electrons is produced in some region intermediate betwee
are first liberated and the surface of the earth. The mean height at wh
an appreciable effect on the direction of propagation of wireless wa
lower limit is by no means sharply defined. It varies considerabl
between night and day.

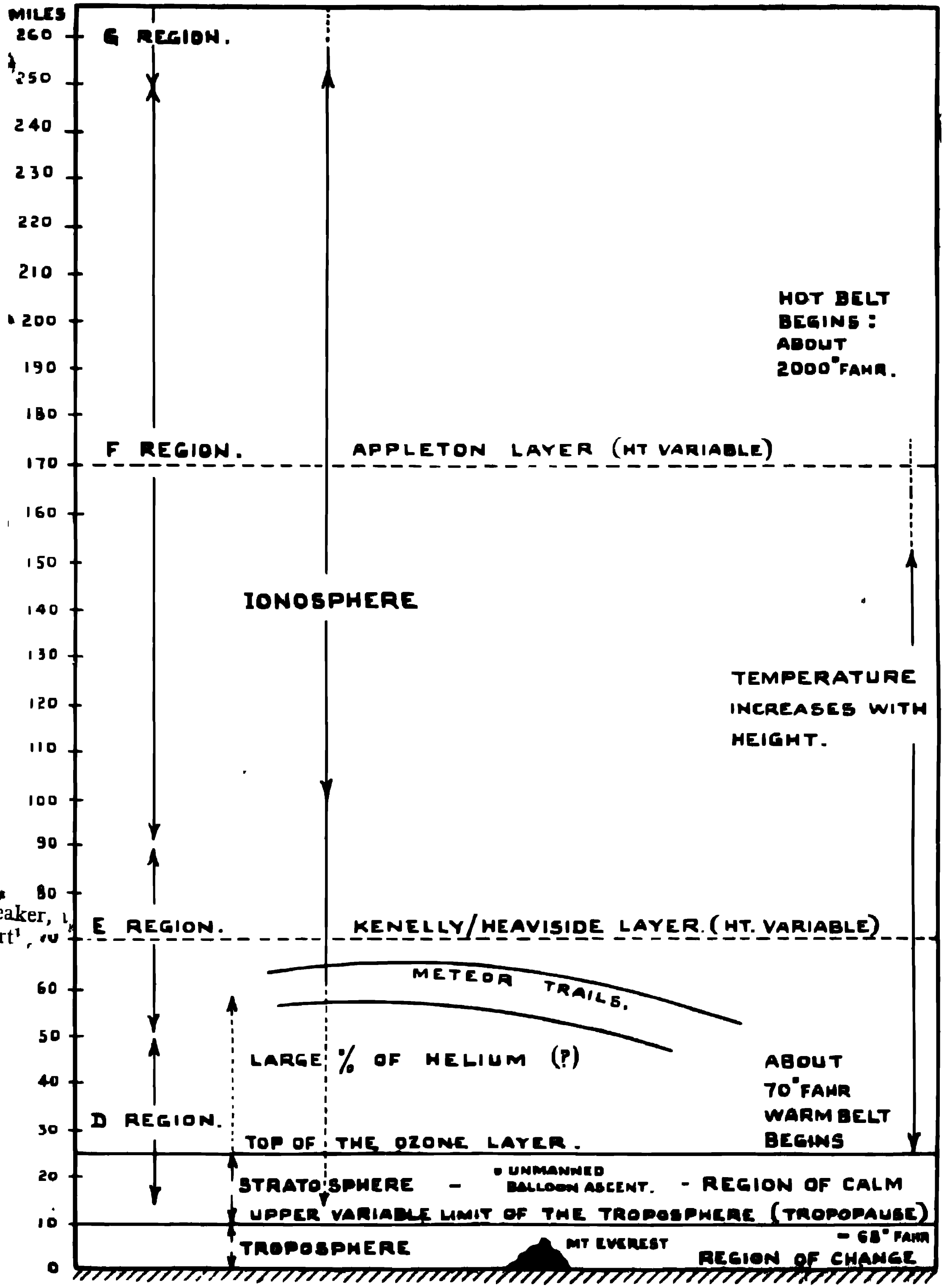

MILES
260
250
240
230
220
210
200
190
180
170
160
150
140
130
120
110
100
90
80
70
60
50
40
30
20
10
0
G REGION.
F REGION.
APPLETON LAYER (HT VARIABLE)
HOT BELT BEGINS: ABOUT 2000° FAHR.
IONOSPHERE
TEMPERATURE INCREASES WITH HEIGHT.
E REGION.
KENELLY/HEAVISIDE LAYER. (HT. VARIABLE)
METEOR TRAILS.
LARGE % OF HELIUM (?)
ABOUT 70° FAHR WARM BELT BEGINS
D REGION.
TOP OF THE OZONE LAYER.
UNMANNED BALLOON ASCENT.
STRATOSPHERE - REGION OF CALM
UPPER VARIABLE LIMIT OF THE TROPOSPHERE (TROPOPAUSE)
- 68° FAHR
TROPOSPHERE
MT EVEREST
REGION OF CHANGE

FIG. 1.

In 1912 Eccles suggested that, after sunset, the ultra-violet radiation ceases to be operative, and in the lower regions, where the density of the atmosphere is greater and collisions between electrons and molecules are more frequent, there is a certain amount of re-combination between electrons and positive ions. The ionisation in these regions therefore decreases between sunset and sunrise, and the effective lower limit of the Heaviside layer moves upwards, rising slowly through a varying distance and sometimes disappearing. The layer rapidly falls—or re-appears—after sunrise and assumes its daylight position.

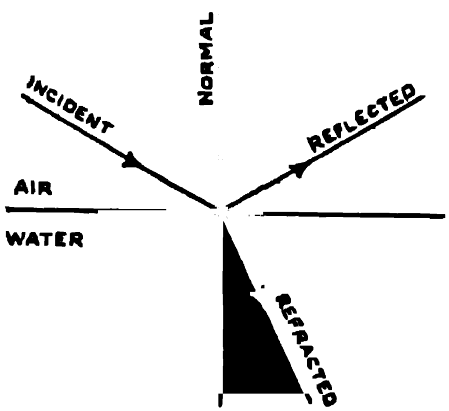
Appleton, in 1925, showed that there was a further densely ionised layer at a greater height than the Kennelly-Heaviside layer; he did this by observing the echo effects from short duration H/F signals projected vertically upwards. The echo signal indicated reflection from a higher layer. The theory has been advanced that the ionisation in the upper air, which is of a more stable and permanent kind, is due to the emission, from the sun, of α particles such as are shot out from radio-active substances. This upper layer comes in the "F" region, and may have a height of anything from 90 to 250 miles, with an average of about 170 miles; this reflecting medium is now called the "APPLETON" layer.

It therefore appears that there are at least three semi-conducting layers of ions widely separated in space, each capable of acting as a reflector of W/T waves of certain frequencies, and all contained within the broad belt now known as the IONOSPHERE. The upper limit of the ionosphere is not known, but the region further away than 250 miles is called the "G" region, and experimenters have recently heard echoes coming back from outer space after intervals of from 3 to 30 seconds, indicating that something akin to reflection must be taking place at distances up to almost 3 million miles from the earth. These echoes can only be obtained on one or two days in each year, and there is much argument as to their real cause.

Fig. 1 shows, diagrammatically, the various media and ionised layers, those of the most importance being in the D, E and F regions.

Before passing to the effects produced by these ionic layers surrounding the earth, it is worth while briefly to consider the basic phenomena of reflection and refraction.

4. **Reflection.**—It is hardly necessary to explain the meaning of reflection. When a billiard ball strikes a cushion its direction of motion normal to the cushion is reversed. In the same way an electromagnetic wave trying to pass from one material medium to another may be turned back at the common surface of the two media. In the case of a light wave arriving obliquely at a point on the surface of a mirror, the reflected wave comes off at the same angle on the other side of the line through the point perpendicular to the mirror at the place where the incident ray touches the surface of separation of the two media, as shown in Fig. 2. Because of this, the reflected image of an object appears to be at the same distance behind the reflecting surface as the object is in front of it. Generally, the wave is only partly reflected at the surface, and a part of its energy passes on into the second medium.



REFLECTION AND REFRACTION.

FIG. 2.

5. **Refraction.**—If a walking-stick is partly immersed in water, it no longer appears to be straight The part under water appears to run in a different direction from that in the air. The light waves, by which the stick is made visible, alter their direction of propagation when they pass from air to water and *vice versa*, and so the stick appears to be bent. The reason for this is that the waves do not travel with the same velocity in the two media. In this case they travel more slowly in water than in air (or rather, more slowly in æther as modified by

ater than in æther as modified by air), and the effect is that the direction of propagation in water bent in towards the "normal," or line drawn at right angles to the air-water surface through the oint where the stick enters the water. The waves are said to be refracted, and the ratio of the elocity of the waves in free æther to the velocity of the waves in a material medium is known as he "refractive index" of the medium. Since the waves travel more slowly in water than in air, he refractive index of water is greater than that of air.

★6. MATHEMATICAL NOTE.—The actual figures are 1·33 for water and 1·00029 for air (at normal pressure and mperature).

Thus
$$n \text{ (refractive index)} = \frac{3 \times 10^8 \text{ metres per sec.,}}{\text{velocity in medium}}$$

$$\text{velocity in medium} = \frac{3 \times 10^8 \text{ metres per sec.}}{n}$$

From the nature of electromagnetic waves we should expect that their velocities in various media would depend on the electrical and magnetic properties of the media, and it can be shown that if K is the dielectric constant of a medium and μ is its permeability, the velocity of electromagnetic waves in the medium is $\frac{3 \times 10^8}{\sqrt{\mu K}}$ metres per second. Since μ and K are both taken as unity for a vacuum, this gives the velocity of electromagnetic waves in free æther as 3×10^8 metres per second, as previously stated.

The permeability, μ, is approximately unity, except in the case of ferromagnetic metals, and so the velocity in a given medium may be taken to be $\frac{3 \times 10^8}{\sqrt{K}}$ metres per second.

It follows from this that the refractive index of a medium is equal to the square root of its dielectric constant,

i.e.,
$$n = \sqrt{K}.$$

This formula is only strictly correct for a perfect insulator. For poor insulators and conductors it must be modified to take account of their conductivity.

The dielectric "constant" of an insulator is not constant, but varies with the frequency of the alternating P.D. applied across it. Thus, a material medium has a different K for electro-magnetic waves of different frequencies, and therefore a different refractive index. The dielectric constant of distilled water at normal temperatures, for example, as deduced from steady potential measurements, is about 80. For alternating potentials of the frequency of light waves it is in the neighbourhood of 2, ($n^2 = 1 \cdot 7$). In other words, the velocity of an electromagnetic wave in any mediumexcept the free æther depends on its frequency. This can be readily seen in the case of light waves by examining the light from the sun through a glass plate.

7. **Dispersion.**—Sunlight consists of a very large number of light vibrations of different frequencies. These all travel with the same velocity through free space, and with sensibly the same velocity through air, owing to its small refractive index, and so combine to give the impression of white light. The differences in velocity of the component vibrations of different frequencies are, however, appreciable in the case of glass. In consequence, these vibrations are refracted through different angles in passing through the glass, and re-appear at different points on the other side of the plate, as shown in Fig. 3. Differences in frequency in the visible range appear to the eye as differences in colour, and so a series of differently-coloured images of the sun will be seen side by side. The white light is said to be resolved into a spectrum. In practice these images will largely overlap each other. To produce a good spectrum, a ray of sunlight entering a dark room through a very narrow slit should be allowed to fall on a glass prism. A series of images of the slit will then be seen, appearing as parallel differently coloured bands of light. The red band is least, and the violet band most, deviated from the direction of the original white light. This phenomenon is called "dispersion."

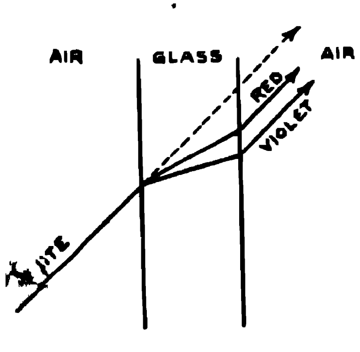


FIG. 3.

8. **Total Internal Reflection.**—It has been seen that a light wave in passing from air to water is refracted towards the normal (Fig. 2). Conversely, a wave from water to air is refracted away from the normal. As the direction of the wave in water becomes more oblique to the normal, the angle made with the normal by the refracted ray in air becomes correspondingly greater, and eventually a point is reached when the wave, on emerging into air, just grazes the common surface, i.e., is at right angles to the normal. If the obliquity of the wave in water is increased further, no wave can then emerge into the air, i.e., the wave is totally internally reflected at the common surface. The angle made by the incident ray with the normal when this occurs is known as the "critical angle."

This phenomenon can only occur when the refracted wave makes a greater angle with the normal than the incident wave, i.e., total reflection can only occur when the electromagnetic waves attempt to pass into a medium in which they would travel more quickly.

9. **Reflection and Refraction in Ionospheric Layers.**—It may be said that the phenomena of reflection and refraction both enter into the explanation of the passage of wireless waves from the transmitter to the receiver. Whenever the indirect ray arrives at the effective surface of separation between an ionised layer and the lower atmosphere, reflection or refraction or both may take place, depending jointly on the angle of incidence, the frequency in use, and the ionic density of the layer.

The effect of ionisation is to make the apparent dielectric constant K less than unity. The result of this is that the higher parts of the wave front in the incident wave will travel faster than the lower parts ; the edge of the wave front where the density of ions or electrons is greatest, will advance faster than the rest of the front, and cause the wave to follow a curved path in which the bending is "away from" the more densely ionised region. Mathematically, this bending phenomenon necessitates the distinction between PHASE velocity and GROUP velocity. The upper part of the wave front, which is being bent downwards, travels faster than the lower part, and is said to possess a greater phase velocity. The velocity of the whole wave front is called the group velocity. In other words, the individual waves composing different parts of the wave front are travelling faster than the wave train as a whole. It is, therefore, clear that the phase velocity may be greater than the velocity of light, whereas the group velocity can never exceed it. In free æther, these two velocities are the same.

Instead of an abrupt change of direction, therefore, the effect of an ionised region is to cause a gradual bending of the direction of the ray. This may be sufficient to bend the wave direction round until it is travelling parallel to the earth's surface, and eventually to direct it downwards so that it again returns to the earth at a distance from the transmitter. The final bending downwards may be due to a process resembling that of total internal reflection.

For a given angle of incidence and mean ionic density, the frequency determines the actual amount of bending that takes place, since the change in phase velocity in different parts of an ionised region depends on the frequency. L/F waves will be bent downwards much more quickly than H/F ones, and it has been calculated that the density of the E region reflecting layer is sufficient to bend back all waves of frequency up to about 750 kc/s., including those incident on the layer vertically Waves of frequency greater than 750 kc/s. break through the E layer and reach the F layer where the ionic density is considerably greater. Even there, however, the density is not sufficient to bend the waves of frequency greater than about 35,000 kc/s., and these break through the layer and are lost in outer space.

It appears probable that most L/F waves are reflected by an ionic layer in the D region. The occasional reception of V.H/F waves at great distances has been attributed to reflection at a low E region of anomalous intensity (paragraph 10).

Fig. 4 summarises, diagrammatically, the reflecting properties of the ionosphere. For simplicity, it is customary in all cases to assume the process of gradual bending to be replaced by one of simple reflection, in which case the height of the equivalent reflecting surface is considerably greater than the actual height to which the waves penetrate. This is shown in Fig. 5.

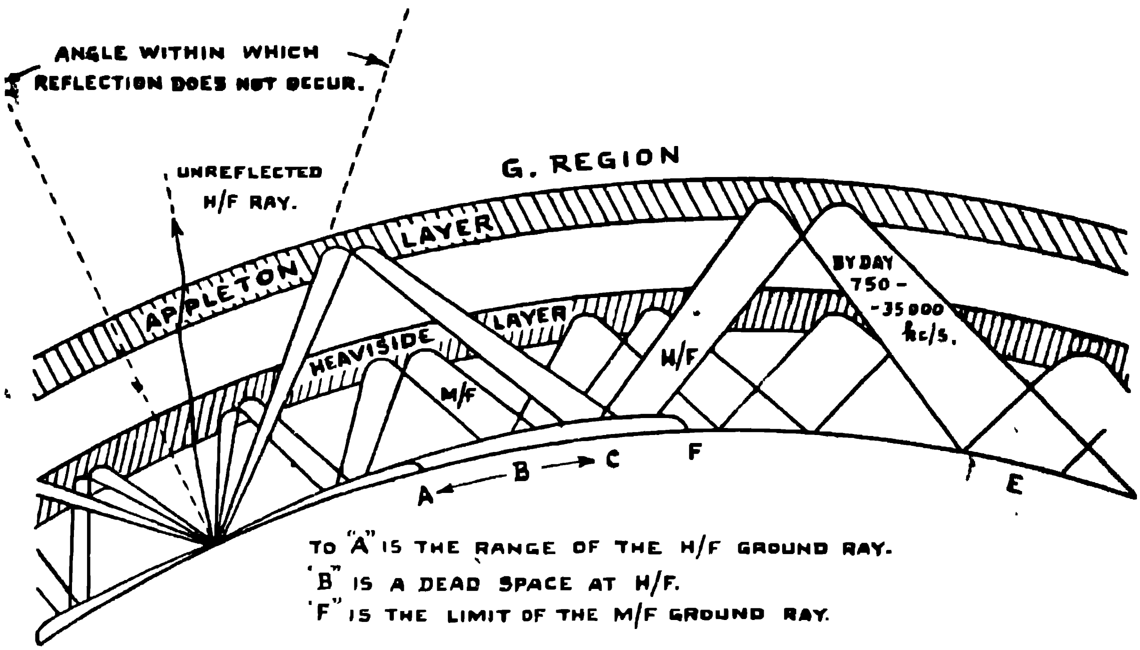


FIG 4

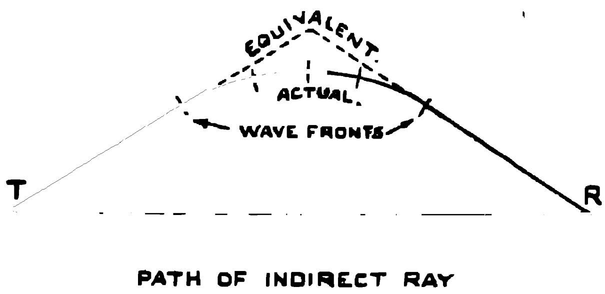


FIG 5

10. **Skip Distance—Day and Night Frequencies.**—For any given frequency there is one direction which the ascending ray may make with the earth's surface at which it is most quickly bent round in the reflecting layer. The point at which this wave reaches the earth is therefore the nearest point to the transmitter at which signals can be received by means of waves returning from the upper atmosphere. The distance from the transmitter to this point (which is, of course, not a point, but a line round the transmitter on the earth's surface, and would be a circle if the radiation and the effect of the reflecting layer were the same on all bearings) is known as the SKIP DISTANCE. Within this distance, the only signals that can be received are those due to the direct wave travelling along the earth's surface. Fig. 4 illustrates this and other matters, further explained below.

In the case of H/F waves, the range of the earth-bound ray may be considerably less than the skip distance, and there appears a zone of silence in which no signals are received. This is shown diagrammatically in Fig. 4. The diagrams also show that the downcoming wave may be reflected at the earth's surface, travel up again to the Ionospheric Layer, be bent round and return

to earth still further away, giving rise to other zones of silence. The regions where no signals are received are generally called DEAD SPACES, being known in order as the first dead space, second dead space, and so on. Those parts of the earth's surface, where signals due to the refracted radiation can be received, are called zones of reception. It will be seen, as a matter of simple geometry, that the width of the second dead space is less than that of the first dead space, and so on, and that the width of the zones of reception increases correspondingly.

Fig. 4 shows that there is a certain angle of incidence within which reflection of (say) H/F rays does not take place. One ray, incident more nearly vertically, is shown penetrating the Heaviside and the Appleton layers before passing out into space. This incident ray travels at an angle less than the critical one, and, accordingly, total internal reflection does not take place. The critical angle becomes greater as the frequency increases, and at V.H/F, a beam leaving the earth at an angle of a few degrees to the horizontal may never suffer reflection anywhere.

Summarising—the skip distance for low frequencies, after reflection by the D layer, is so small that it is far smaller than the ground ray range. Moreover, the loss of energy on reflection is so great that at a distance from the transmitter greater than the ground ray range, the rays have been reflected between the earth and the D layer so many times that their strength is practically nil and they do not increase the range of the transmitter. The medium frequencies may have a ground ray range of up to 1,000 miles, and the indirect ray consists of a series of reflections between the Heaviside-Kennelly layer and the earth. The high frequencies may have a ground ray range of up to a few hundred miles, and the indirect ray is refracted and attenuated in the Heaviside layer before being reflected in the Appleton layer. This process may be repeated several times

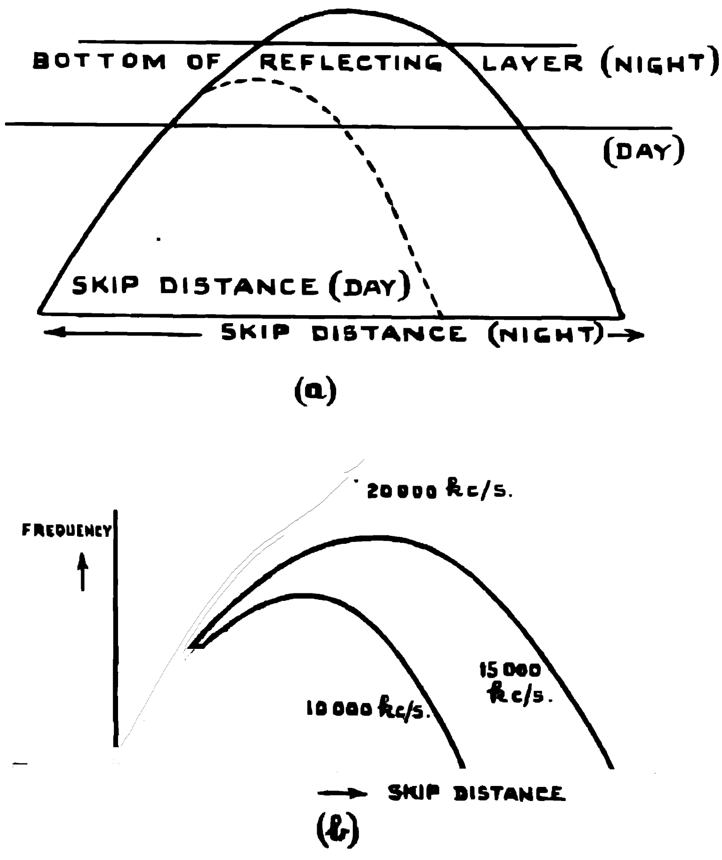
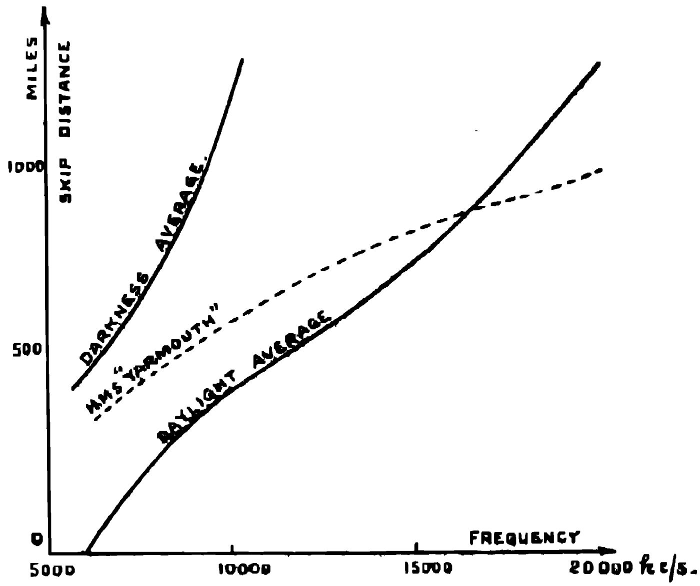


FIG. 6.

and H/F often gives world-wide range by means of the indirect ray. Very high frequencies have a ground ray range of only a few miles, and, in general, the indirect ray passes out into space. These frequencies are capable of providing an excellent service within optical range only; in order to make the horizon as distant as possible, it is necessary to locate the transmitter and receiver as high as possible.



VARIATION OF SKIP DISTANCE WITH FREQUENCY.

FIG. 7.

The effective equivalent height of the layer is not the same by day as by night and, with reference to Fig. 4, it follows that the actual positions of the areas C, E and F alter, and a ship that is within the range of the "first skip" by day may be in the "first dead space" by night. This is further illustrated in Fig. 6 (*a*), and Fig. 6 (*b*) indicates the remedy to be applied Waves of lower frequency are more quickly bent round than waves of higher frequency, and, accordingly, different frequencies must be used for day and night working. For example, in one case, two related frequencies are 6,450 kc/s by day and 5,000 kc/s. by night. The increase in skip distance, due to an increase in the vertical height of the layer by night, is cancelled by the use of the lower frequency. In many cases, it is necessary to have more than two frequencies if a 24-hour service is to be maintained between two points This is one of the matters that somewhat complicates the B.B.C. Empire H/F beam transmissions

The highest daytime frequency of any use for long distant telegraphy is of the order of 25,000 kc/s. (12 metres), and it is probable that this figure is too high unless one is somewhere near the optimum period in the sun spot cycle (*see below*).

To illustrate the above remarks, the variation of skip distance with frequency, at high frequencies, is shown in Fig. 7, the two full lines corresponding to paths between transmitter and receiver lying altogether in daylight and darkness respectively.

The increase of skip distance with frequency and the much larger skip distances at night should be noted. The shapes and relative positions of the curves are probably fairly accurate, but the actual figures should be treated with reserve, since the experimental evidence is rather conflicting.

The values given are average values over a year. In summer the skip distances are less than in winter. Some figures obtained in H.M.S. "Yarmouth" on a cruise to Hong Kong and back in 1925 are given in the following table, and are also shown in Fig. 7 by the dotted line.

f (kc/s.).	Skip distance (miles).	Range of earthbound component (miles).	Width of first dead space (miles).	Width of first zone of reception (miles).	First overlapping zones.
25,000	1,100	30	1,070	800	2nd and 3rd.
12,000	720	80	640	440	2nd and 3rd.
8,500	520	170	350	240	3rd and 4th.
6,250	350	350	Nil	60	No overlapping up to 6,000 miles.

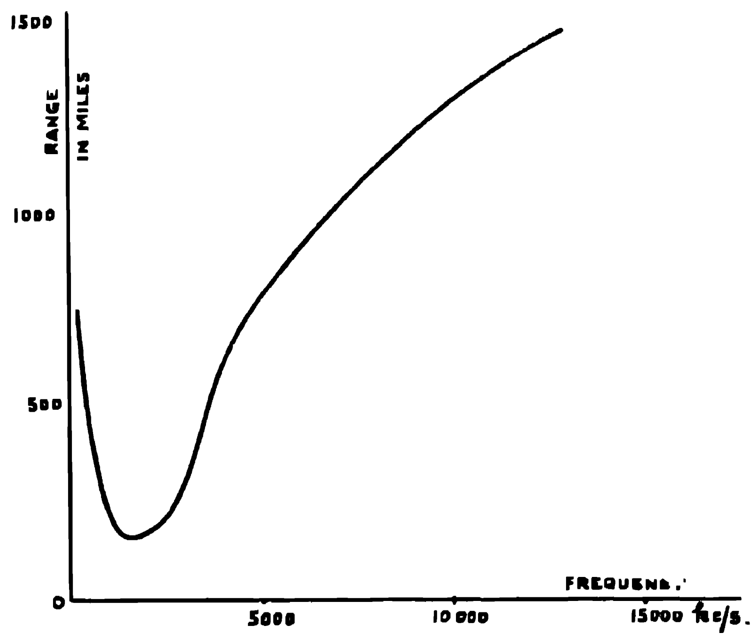
It is not strictly correct to say that no signals can be received in dead spaces, but there the signal intensity is very weak compared with that in the zones of reception (less than one ten-thousandth). Such signals as are obtained are attributed to "scattering" of the indirect waves in the ionospheric layer. Instead of being gradually bent round, part of the stream of energy is scattered in various directions, and so may descend to earth at any angle. A similar phenomenon in the case of light waves is responsible for the blue colour of the sky during the day. If the earth possessed no atmosphere, the only sunlight waves reaching it would be from the direction of the sun, and the sun would appear as a bright disc in a black sky (apart from the stars). But the atmosphere scatters the sunlight, and it appears to reach the earth from all directions, giving rise to the familiar blue colour of the sky during the day. An imaginary observer in the E region would see a black sky.

Since the ionisation of the E and F layers depends largely upon the sun, the ionisation intensity will vary with the seasons—Summer and Winter—as well as with the change from day to night. Some of the British investigators have recently shown that, under Summer conditions, extra ionisation often forms a specially intense E region, of such an ionic density that it may produce ANOMALOUS REFLECTION OF ALL H/F AND V.H/F WAVES, with consequent alteration of skip distance and dislocation of H/F traffic. There is a further slow period of change of about 11½ years, which seems to be determined by the variety of the number of the sunspots appearing upon the solar disc. These spots are supposed to represent enormous cyclonic upheavals, during which the stream of electrons poured out is enormously increased. In 1934, the sunspot cycle was at its minimum, with corresponding minimum effect on the ionisation of the layers. Wave frequencies for satisfactory communication to distant parts may become slightly higher, up to the sunspot maximum period in 1939.

Considerable fluctuations in signal strength will occur when a signal passes from a transmitter in daylight to a receiver in darkness and *vice versa*. In that part of the path where sunrise or sunset is occurring, the density of ionisation in the lower part of the layer is changing rapidly, and evidence of this is to be found in the large changes of signal strength at the receiving station.

11. **Attenuation of Indirect Rays.**—Attenuation will necessarily occur in the indirect rays since the atmosphere is not a perfect dielectric. The great ranges attained with small power on H/F waves show, however, that their attenuation in the atmosphere is much less than that in the surface of the earth. For H/F waves the losses in the lower atmosphere are negligible, though they are a considerable cause of attenuation in the case of very L/F indirect waves. The important attenuation of indirect H/F waves takes place while they are travelling through the ionosphere. This attenuation is roughly proportional to the density of free electrons. Thus it is greater by day than by night, and so night signals are stronger than day signals. It also varies inversely as the square of the frequency. Hence, other things being equal, the higher the frequency the less

the attenuation. This helps to fix a LOWER LIMIT to the frequencies which can be successfully used for long-distance transmission. The limits appear to be about 8,000 kc/s. for regular day transmission and 4,000 kc/s. for regular night transmission. Waves of low frequency suffer such great losses in transmission that they are useless for long distances unless exceptionally high power is developed in the transmitting aerial. These losses occur partly in the ionosphere, and partly at the earth's surface when multiple reflection takes place. The great attenuation of H/F waves at the earth's surface has previously been mentioned when considering the short range of the direct ray. In addition, unless the surface is perfectly plane, the wave will not be regularly reflected, and the energy will be scattered in various directions, so that only a small part is available for any particular transmission under consideration. This, in itself, almost limits long-distance transmission to rays which have only undergone one or two reflections at most at the earth's surface. The energy of these lower frequency waves cannot, of course, penetrate the Heaviside Layer, and so must be dissipated in some such manner as that described above before the wave reaches great distances from the transmitter. In the case of transmission over shorter distances at these frequencies, or of long-distance transmission at higher frequencies, the strength of a received signal at a particular point may thus depend very greatly on the conditions at a point half-way between it and the transmitter, e.g., a jungle at a distance of 2,000 miles may render reception impossible at 4,000 miles on the same bearing.



VARIATION OF DAYLIGHT RANGE WITH FREQUENCY.

FIG. 8.

12. **Electron Resonance—Critical Frequency.**—Since in typical low frequency transmission the attenuation increases with the frequency, and the reverse is the case in long range high frequency transmission, it would be expected that at some intermediate frequency minimum ranges would be obtained. This is actually the case, as shown in Fig. 8 by the curve of daylight range against frequency for constant energy radiation from the aerial. It will be seen that there is a distinct falling-off in range at frequencies between 1,000 and 2,000 kc/s., much larger, indeed, than would

be expected from the considerations advanced above. Frequencies in this range are not low enough to operate satisfactorily using the ground ray, and not high enough to maintain reliable contact using the indirect ray.

An attempt has been made to explain this large attenuation by showing that at such frequencies the combined results of the fields of the wave and the earth's magnetic field is to produce a resonance effect in the motion of the free electrons in the ionospheric layer. It can be shown that when the frequency is low, the oscillating electrons follow elliptical paths. As the frequency increases the electron orbits become circular in shape when the frequency is in the neighbourhood of 1,400 kc/s.; this is the resonant condition and the orbits then tend to increase in diameter until collisions occur with the gas molecules. These collisions, and the resonant state, produce an absorption of energy resulting in the attenuation of the wave If the frequency is increased above this critical frequency, it can be shown that the electron orbits then become elliptical in shape, the former minor axis then becoming the major one. At night time the ionisation intensity is less and the layer rises. Since the electron density and the gas density are both relatively small, the chances of collision decrease by night and attenuation of the wave due to this cause is correspondingly smaller.

13. **Polarisation and Phase Changes.**—Changes in the nature of the polarisation of the indirect ray in its path through a reflecting layer are mainly attributed to the effect of the earth's magnetic field.

When the ray enters the layer it is split into two oppositely rotating components. When these combine, after emerging from the layer, rotation of the plane of polarisation will be produced and this will give rise to certain difficulties in direction finding, which are referred to in Section "T."

More generally, the emerging down-coming ray is found to be elliptically polarised. It is this that accounts for the fact that in long distance H/F transmission equally good reception is obtained using either horizontal or vertical receiving aerials and is independent of the nature of the transmitting aerial. This is by no means the case, however, if the direct ray from an H/F transmitter is being used.

As in the case of light waves, the coefficient of reflection of horizontally polarised waves (the electric vector parallel to the ground) is greater than if the electric field vector is vertical. Moreover, there is a certain critical angle of incidence, for vertically polarised waves, after which a change of phase of 180° occurs in the reflected wave; this critical angle is known as Brewster's angle (Section R.40).

14. **Fading.**—This is the name given to the occurrence of fluctuations in signal strength at the receiver. The period of the variation may be long or short. It is specially noticeable at night for frequencies between 500 and 1,000 kc/s. at distances of the order of 100 to 1,000 miles from the transmitter. It is violent at night for frequencies between 1,500 and 5,000 kc/s. at distances from the transmitter of 5 to 300 miles. At higher frequencies it occurs mainly at the edges of the zones of reception.

Fading may be due to various causes. It is reasonable to suppose that the edges of zones of reception are not rigidly delimited but subject to small rapid fluctuation, corresponding to similar fluctuations in the effective height of the reflecting layer. A variation in the ionic density would produce such a result. When the frequency and distance from the transmitter are such that both the direct and indirect rays are received with comparable intensities, the phenomenon known in optics as "interference" will result; when the difference in path length is such that there is a phase difference of 180° between the energy received along the two paths, a nil effect will be produced. Conversely, of course, when the difference in path length is a whole number of wavelengths, the direct and indirect rays will reinforce each other. The cathode ray direction finder is capable of giving visual indication of the arrival of waves at the receiver from different directions.

It is usually observed that the different frequency components of a modulated wave fade in a manner in which each is independent of the other. This has been termed "selective fading," and the effect is that the received signal is not the same as the transmitted one and "quality distortion

is the result. The use of I.C.W. at H/F is, therefore, a means of combating fading in telegraphic work.

15. **M/F Broadcast Transmissions—Service Area.**—One of the aims of the radio engineer in designing a broadcasting service is to provide a field strength giving a satisfactory signal/noise ratio over a given "service area" around the station. In this connection it may be noted that a committee o the I.E.E. recently decided that there should be a difference of at least 40 decibels between the strength of the signal (or wanted) field and the interfering (or unwanted) field. With reference to this signal strength criterion, it is possible to recognise certain areas with distinctive properties surrounding any station. Near the station we have an area in which the ground ray preponderates, having a signal strength much greater than that of all interference. In this "service area" the signal strength may be relied upon to be satisfactory at all times. Surrounding this area is one in which the signal/noise ratio is unsatisfactory, although the ground wave may still be much stronger than the sky wave. Further away, in general there is a region in which the sky wave is almost equal in strength to the ground wave. This produces fading and quality distortion. Still further away, the whole area is surrounded by the region which is served by the sky wave only; the signal strength may never be very great and will be subject to large seasonal and diurnal variations, being at its best at night time and in the winter. Fig. 9 illustrates diagrammatically the meaning of these remarks.

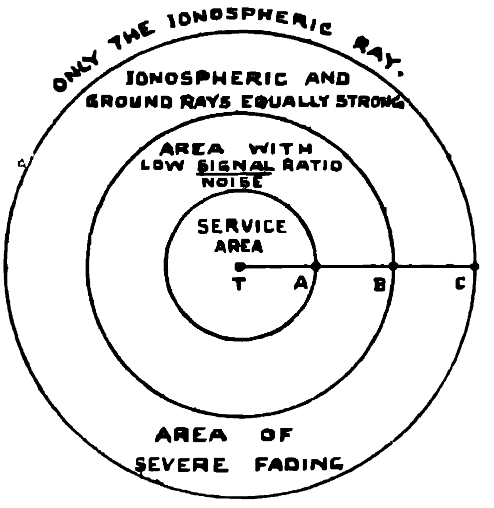


FIG. 9.

It is therefore clear that the most useful service area of a broadcasting station is the area in which the ground ray is much stronger than the sky ray. Under the best conditions, the service area will be as large as possible when its edge extends to the point B of Fig. 9, i.e., when the area immediately surrounding the service area does not exist; in practice, this will seldom take place. The service in the fading belt can only be improved by reducing the proportion of the sky wave; if the indirect ray is as strong as the direct ray, an increase in the power of the transmitter will merely increase the signal strength of both rays and not alter their relative amplitudes. The cure for fading would be the complete elimination of the sky wave; this may not be possible, but the new anti-fading aerials provide a concentration of energy in low angle radiation, greatly reducing the sky wave and simultaneously increasing the ground wave (R.34 and R.40). All areas in Fig. 9 are thereby increased. It would appear that if the use of anti-fading aerials becomes at all universal, the possibility of listening to foreign stations on the M/F wave band may be much reduced; the reception of distant stations depends almost entirely upon the ionospheric ray. It is, however, considered that this is not too high a price for the broadcast listener to pay for the reduction of fading troubles on home stations.

16. **Choice of Frequency for H/F Communication.**—It may be said that the best frequency to use is the highest one which will give a signal at a distant point. With this condition it is usually safe to say that the signal passes to the receiver in the fewest steps and, therefore, with the minimum attenuation. Reference has already been made in paragraphs 10 and 11 to the upper and lower limits of frequency for satisfactory service, and it has also been made clear that, in general, various frequencies may be necessary to maintain a 24-hours service.

17. **Multiple Signals—Echoes.**—Fig. 10 shows that signals may be propagated to a distant

point by paths of different length. In the case of telegraphy, this means that the time of arrival of the various signals corresponding to (say) a Morse dot will not be identical; one signal follows another and produces blurring or double signals, an effect which is specially serious when working at high speed. The presence of multiple signals puts a limit to the speed of high speed working. Sometimes a signal echo is received at a considerable time after the receipt of the original. It is thought that short echoes may be due to a signal arriving at a receiving station having been the longer way round on the great circle path. Very long echoes, up to several minutes, have been variously explained; they may be due to the signal making various circuits of the earth before emerging, or, alternatively, the signal may have journeyed to some distant reflecting layer in outer space.

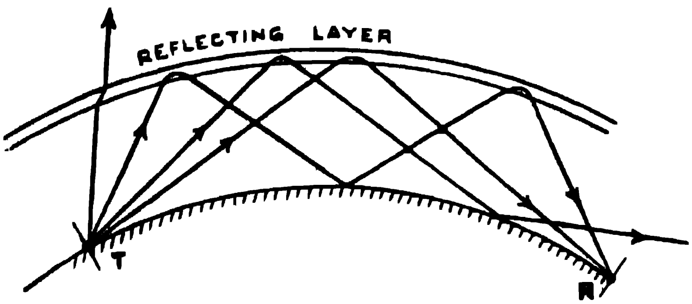


FIG. 10.

Very short duration wireless signals, pulses lasting about 10^{-4} seconds, can be made to produce echoes from the various ionospheric layers at a measurable interval of time after the transmission of the original pulse. This principle is the basis of the radio echo method of investigating the upper atmosphere. The measurement of the very small intervals of time involved is made possible by the use of the cathode ray oscillograph, by means of which it is easy to measure time intervals to an accuracy of a millionth of a second. Much research work is being conducted along these lines by investigators in various parts of the world. Colwell and Friend, working in America, have recently used pulse signals of only one hundred thousandth of a second's duration, by which means they have been able to detect a path difference of as little as 3 kilometres between the "ground" and the "echo" signals. Similar work in this country is being conducted by a British team of investigators, some of them working under the auspices of the Radio Research Board of the Department of Scientific and Industrial Research.

18. **Man-Made Noise—Atmospherics (Statics).**—It is well known that in spite of perfection in the propagation conditions and at the transmitter, received signals may be almost unintelligible due to noise and static, the various clicks, bangs, rumbles and crashes which often provide the background of the signal. This noise, in general, is classified into "man-made noise" and "statics," and is the noise which may disappear when the aerial is disconnected from the receiver.

Man-made noises may be due to electrical machinery of various kinds, ignition systems of cars, motors, etc. Clearly, interference of this type is capable of suppression, and it is probable that in the not too distant future, no new electrical apparatus will be put into use which has not been previously certified as "interference free." This has recently been the subject of a special report by a committee of the I.E.E. Where suppression at the source is not undertaken, some improvement in reception is usually noticed when the vertical down-lead of the receiving aerial is screened; this is due to the fact that man-made static consists of H/F signals which are, in the main, vertically polarised.

Statics, or atmospherics, are radio waves produced by natural causes, of definite but very irregular wave form and, generally, of short duration (about 1/500 of a second). The peak voltage produced by a static signal often amounts to as much as 1·5 volts, the energy level being very great at the lower radio frequencies, decreasing gradually as the frequency becomes greater, and becoming quite small at V.H/F. The very complex wave form of an atmospheric makes it equivalent to a very large number of simple sinoidal wave forms, with the well-known result that statics appear to be untunable noises covering up a wide frequency range. Up to the present time, there have

been two principal ways by which the effect of static at the receiver has been minimised in order to increase the signal/noise ratio. Briefly, these involve

(a) receivers with high selectivity, and
(b) the use of directional reception.

With constant frequency transmitters, for purposes of telegraphy, the receiver may be made very highly selective, and consequently able to reject many of the interfering frequencies constituting a static noise. Under good conditions it is possible to work with receivers having a band pass width of only 50 cycles. The case, of course, is different for radio telephony, where a much wider band of frequencies must be passed, especially if high quality speech or music has to be received.

Although static noises are liable to come from any random direction, each one individually is directional in nature. For this reason it has been found that much improvement of the signal/noise ratio is attained by the use of an arrangement of aerials for directional reception. Some of the better known arrays for directional reception are described in Section " R."

Quite recently, a new method of static suppression has been suggested, based upon the commonly observed high value of the amplitude of the disturbance. In principle, it resembles the action of the various well-known A.V.C. circuits ; the high amplitude of the static is used to trigger a valve which provides a cut-off potential at the grid of one of the A/F stages and preventing either the signal or the disturbance from being heard. The receiver is muted for a time corresponding to the duration of the static, but, since the cut-off time is very short, it does not affect the apparent continuity of the signals. It appears possible that some device working on this principle may provide a contribution towards the solution of the static problem.

19. **The Luxembourg Effect.**—While listening to a distant medium frequency broadcasting station, it is sometimes found that there is a continuous background of some more powerful station, usually of lower frequency. This effect was observed when R/T stations first began to use high power. In particular, English listeners to Radio Luxembourg (230 kc/s.) noticed that its transmission was accompanied by a background of that from Radio Paris (182 kc/s.) ; the phenomenon was called the " Luxembourg effect," although many other similar " effects " have since been observed in connection with other stations.

The phenomenon has been the object of much investigation and an explanation of it was first given by two Australian physicists, Bailey and Martyn ; they showed how a powerful L/F station could affect the ionosphere in such a way that any other waves reflected from the affected region would acquire a modulation from the unwanted powerful station. To quote Professor Appleton—" the ionised region appears, as it were, to have the programme of the powerful station stored up in it, and passes a little of it on to any other wave which it reflects." From the investigation it was apparent that this type of interference would be most marked when—

(a) the two transmitting stations were approximately on the same great circle bearing from the receiver, and

(b) the interfering station was geographically between the receiver and the wanted transmitting station.

The above results are those commonly observed in practice. The only cure for this unfortunate mixing of transmissions would appear to be a reduction of the proportion of the sky wave.

A partial explanation of the phenomenon can be given in the following terms :—

Under the action of a powerful electric field, such as that above a powerful transmitting station, the motion of the electrons does not follow a linear law. It is possible to make use of this non-linear characteristic, in the way in which one does when employing " grid modulation." By a process resembling the latter, the waves of the distant station passing through the area of the powerful electric field have a modulation from the latter impressed on them.

EXAMINATION QUESTIONS ON " PROPAGATION."

1.—(*a*) Describe the nature and properties of the Heaviside layer, with particular reference to the effect it has on—(i) the path, and (ii) the attenuation of indirect rays.

(*b*) What are the factors influencing the attenuation of the direct ray from a transmitter?

(*c*) Explain the reasons for the fading of H/F signals which may occur at a receiver.

(Qualifying for Lt. (S), 1932.)

2.—(*a*) Describe and account for the ionised state of the upper atmosphere, and its daily and seasonal variations.

(*b*) Discuss the factors influencing the choice of a frequency (high) for long distance communication, and show how a knowledge of the conditions existing along the path of the signal is of assistance in selecting a suitable one.

(*c*) Account for the existence of a critical H/F band unsuitable for long distance communication.

(Qualifying for Lt. (S), 1933.)

3. Give a short account of the propagation of wireless waves with particular reference to skip distance, attenuation, day and night frequencies, dead spaces, and polarisation.

(Qualifying Warrant Tel., 1931.)

4. Explain the causes of fading in the reception of short and medium waves. What methods are adopted in practice to minimise the effects of fading?

(C. & G. Final, 1933.)

5. What is the part played by the Heaviside layer in the propagation of short waves? Why is it necessary, in short wave communication over long distances, to use longer waves during the night than during the day?

(C. & G. Final, 1934.)

6. What are the causes of fading :—

(*a*) In the case of medium wave stations, within range of the ground ray;

(*b*) In the case of short wave stations beyond the range of the ground ray?

What methods are adopted in practice to combat fading?

(C. & G. Final, 1935.)

AERIALS, FEEDERS, DIRECTIONAL ARRAYS.

TRANSMITTING SYSTEMS.

1. **The Basic Principles of Radiation.**—To appreciate the function and properties of an aerial system, it is essential to have some idea of the way in which a portion of the energy may become detached from an oscillatory circuit, and be radiated away into space. The accurate mathematical analysis was first produced by Clerk Maxwell in 1864 from the general laws of magnetism and electricity, but that is beyond the scope of this book. In spite of the mathematical difficulties, by making certain very reasonable assumptions it can be shown that DETACHMENT OF ENERGY MUST OCCUR WHENEVER THE CURRENT IN A CIRCUIT CHANGES, *i.e.*, whenever an electron has its velocity altered by the action of an accelerating force. It must be pointed out, however, that any attempt to produce a mechanical picture of the radiation process must necessarily be somewhat crude, and in some ways even misleading.

Consider the simple type of oscillatory circuit represented in Fig. 1 (*a*), consisting of a condenser whose plates are connected by a vertical wire which has a certain amount of self-inductance (effectively an aerial circuit) An alternator is included in the circuit to represent a source of

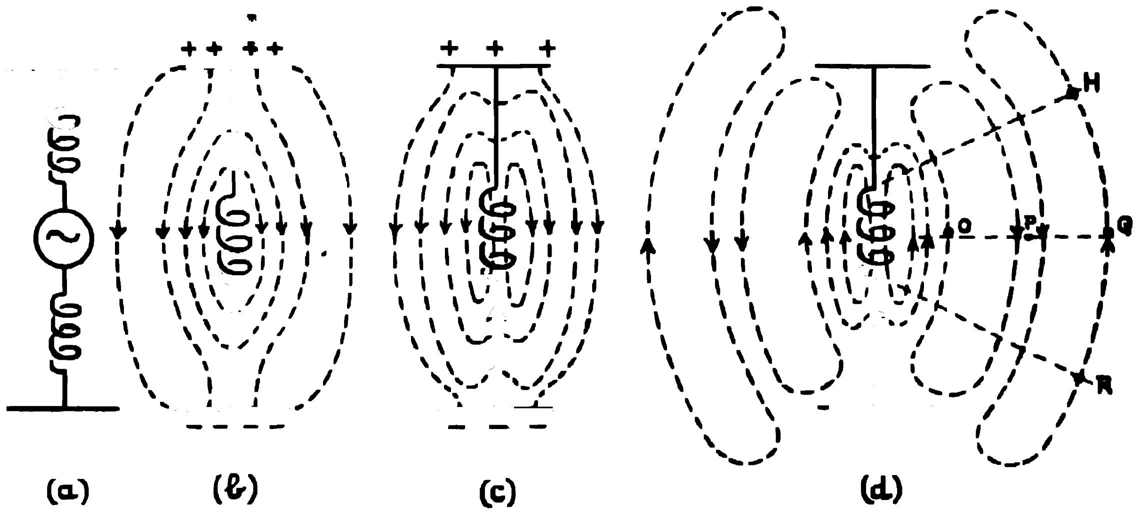


FIG. 1.

alternating E.M.F. of "radio-frequency," the frequency of which will usually be that to which the circuit is resonant: the whole arrangement is sometimes called a **dipole** or Hertzian oscillator.

A dipole may be defined as "any symmetrical aerial the two ends of which are at opposite potential with reference to the central earthy point."

Consider the distribution of the electric field around such an aerial, and the changes in the field during one cycle of the applied E.M.F.

We will consider the cycle to start when the condenser is charged to its maximum P.D., so that the top plate is positive with regard to the bottom plate, and the current in the wire is zero. At this instant we may regard the field in the vicinity of the aerial as being entirely electric, and lines of electric stress to be connecting each positive charge on the upper plate to its "opposite number," a negative charge on the lower plate, as in Fig. 1 (*b*).

When the moment of maximum P.D. has passed, electrons will start to flow upwards, the flow constituting a current of electricity. The electric field starts to collapse, and this effect may be represented as in Fig. 1 (*c*), in which the ends of the lines of force are shown coming together along

the wire. Due to the familiar property of electric inertia, summarised by Lenz's and Faraday's laws, the current continues to flow after the potential difference across the condenser is reduced to zero, and in so doing starts to charge up the condenser in the opposite direction, giving rise to new lines of force in the reverse direction to the previous field.

Now, if we regard the collapse of the initial field as lagging a little on the changes in potential which cause it to take place, then it is clear that the new electric field starts to build up before the first one has disappeared. The first disturbance is then forced outwards in the form of closed loops by the new electric field [Fig. 1 (*d*)], for the direction of the lines in the inner surface of the first and the outer surface of the second are the same, and, accordingly, mutual repulsion takes place.

In this connection it may be observed that, since such a system in free space radiates in three dimensions, a much more correct picture of the process is obtained by visualising an expanding soap bubble having the aerial at its centre; as the radius of the bubble grows, other concentric ones are produced within it at regular time intervals.

It could be added here, that it is *essential* to regard the collapse of the initial field as lagging on the change which causes it, unless one can accept the idea of—

(*a*) the action of a force at a distance, without the help of an intervening medium, or

(*b*) an infinite speed of motion of the flux lines.

In addition to the electric field, the changes in which are represented in Fig. 1, we must regard the circuit as being surrounded by rings of changing magnetic stress, the intensity of which at any point will vary with the current strength, and whose direction alternates.

In the immediate neighbourhood of the current carrying wire, the magnetic lines of force are in the horizontal plane, at right angles in space to the electric flux; it is here reasoned that this essential space quadrature is maintained in the radiation field. In picturing these rings of magnetic stress, the reader should not be disturbed by the coil inductances of Fig. 1, which are only there to represent the lumped inductance of the aerial just as the plates represent lumped capacity.

The electric and magnetic INDUCTIVE FIELDS ARE IN BOTH TIME AND SPACE QUADRATURE, that is to say, they are 90° out of phase in time and at right angles to each other in space, and if an effect at a distance did not lag on the cause producing it in the way referred to above, the energy would simply oscillate from being entirely *potential* when the condenser is fully charged, to being entirely *kinetic* when the current is a maximum, the change taking place without any energy losses. In that case we could then equate the energy in a charged condenser ($\frac{1}{2}CV^2$) to the energy in the magnetic field of an inductance ($\frac{1}{2}LI^2$).

Further examination of the process, shown diagrammatically in Fig. 1, makes it clear that the magnetic field is only present when the electric field is changing; with stationary electric flux there is no associated magnetic field [Fig. 1 (*b*)]. It would therefore appear that any *changing* or moving electric field *must* have a magnetic field associated with it. It is logical to extend this idea to include the case of the loops of the electric field which become detached and RADIATED away. At this point there appears an important and essential difference between RADIATION and INDUCTION, which is not brought out by the necessarily crude diagrams of Fig. 1. The *movement* of such a closed loop becomes contingent on the association of the electric and magnetic fields. If one has a maximum value then so must the other, if the one is zero then so must the other be; the electric and magnetic fields cannot exist separately, and become simply different ways of expressing the fact that energy is being transmitted by an electro-magnetic wave. In more precise terms, this means that the RADIATIVE FIELDS ARE IN TIME PHASE WITH EATH OTHER ALTHOUGH AT RIGHT ANGLES IN SPACE.

Fig. 2 represents the relative directions of the electric field, the magnetic field, and the direction of propagation. The oscillating electric field is shown by the vertical vector OZ, and the oscillating magnetic field by the horizontal vector OX, which is really perpendicular to the plane of the paper. The direction of propagation of the wave, OY, is at right angles to the plane of OX and OZ, *i.e.*, at right angles to the " wave front."

This diagram also makes clear what happens when an E.M. wave is reflected. The simplest case of reflection is when a wave arrives at the surface of another medium, when the direction of propagation is perpendicular to the surface. The direction of propagation is then completely reversed and the wave travels back along its original path. It suffers a phase change of 180°. From Fig. 2 it will be seen that the electric and magnetic vectors are arranged with respect to the direction of propagation, like the fore-finger, middle finger, and thumb of the right hand extended as in Fleming's rule and, as in that case, reversing any two of the three quantities together does not alter the direction of the third. Thus, when the direction of propagation is reversed, the direction of either the electric field or the magnetic field will be reversed but not both. Usually it is the electric field which reverses in direction, or, in other words, changes its phase by 180° at a reflecting surface; the magnetic field is unaltered.

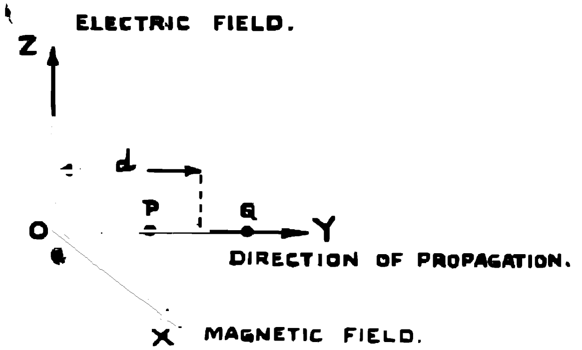


FIG. 2.

2. **Velocity of Propagation.**—If we measure any electrical quantity, such as the capacity of a condenser, in electrostatic units and also in electromagnetic units, the ratio of the two involves a constant "c," which can be shown both to have the dimensions of a velocity and to be equal numerically to the velocity of light in free space. In 1864 Maxwell's purely mathematical reasoning enabled him to predict that changing electric fields and the changing magnetic fields that they produce, *must* give rise to electro-magnetic radiation, consisting of a complementary condition of electric and magnetic fields, whose velocity of propagation through space would be "c" When the measured value of "c" was found to be that of the velocity of light, it was then only natural that scientists should produce the theory that light itself was of electro-magnetic origin, and should try to produce similar radiation by electrical means. In 1888 the successful production of such radiation by Hertz, established the truth of Maxwell's theories, and laid the foundation of Wireless Telegraphy. Much corroborative evidence has since shown that not only light, but also X-rays and γ-rays are of the same nature.

The same theory shows that the amplitude of the electric field $\mathscr{X}$, measured in electrostatic units, is equal numerically to the amplitude of the magnetic field $\mathscr{H}$, measured in electro-magnetic units. When both $\mathscr{X}$ and $\mathscr{H}$ are expressed in the same fundamental units, the relation connecting them is

$$\frac{\mathscr{X}}{\mathscr{H}} = c \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad \text{where } c = 3 \times 10^{10}\,\frac{\text{cms.}}{\text{sec.}}$$

In R.M.S. values the relation is $X = cH$

.... but using particular practical units this formula may be transformed. In E.M.U's and in air, H is expressed in $\frac{\text{lines}}{\text{cm}^2}$; if H is $\frac{1 \text{ line}}{\text{cm}^2}$

then .... $X = 3 \times 10^{10} \times 1$ .... in E.M.U's of field strength,

or .... $X = \dfrac{3 \times 10^{10}}{10^8}$ .... in $\dfrac{\text{volts,}}{\text{cm.}}$

the general formula being .. $X = 300H$ .. in practical units. [*Cf.* Section "T," paragraph 5.]

★3. Relation between E.M.U's and E.S.U's.

Electrostatic Units.—This system is based on the force between two charges given by the formula :—

$$F = \frac{Q_1 Q_2}{Kd^2}$$

Writing this as a " dimensional " equation we have

$$\left[MLT^{-2} \right] = \left[\frac{Q^2}{KL^2} \right]$$

Hence the dimensions of Q are given by

$$\left[Q \right] = \left[K^{\frac{1}{2}} M^{\frac{1}{2}} L^{\frac{3}{2}} T^{-1} \right]$$

The dimensions of electric current are then

$$\left[i \right] = \left[K^{\frac{1}{2}} M^{\frac{1}{2}} L^{\frac{3}{2}} T^{-2} \right] \qquad \text{(1)}$$

Electromagnetic Units.—This system is based on the force between two poles given by the formula

$$F = \frac{m_1 m_2}{\mu d^2}$$

Dimensionally

$$\left[MLT^{-2} \right] = \left[\frac{m^2}{\mu L^2} \right]$$

$$\left[m \right] = \left[\mu^{\frac{1}{2}} M^{\frac{1}{2}} L^{\frac{3}{2}} T^{-1} \right]$$

Field strength (*i.e.*, force per unit pole) at the centre of a circular coil is given by :—

$$H = \frac{2\pi i}{r}$$

or, dimensionally

$$\frac{\left[MLT^{-2} \right]}{\left[\mu^{\frac{1}{2}} M^{\frac{1}{2}} L^{\frac{3}{2}} T^{-1} \right]} = \frac{\left[i \right]}{\left[L \right]}$$

$$\therefore \quad \left[i \right] = \left[\mu^{-\frac{1}{2}} M^{\frac{1}{2}} L^{\frac{1}{2}} T^{-1} \right] \qquad \text{(2)}$$

Now, electric current can have only one set of dimensions, so from (1) and (2)

$$\therefore \quad \left[\mu^{-\frac{1}{2}} M^{\frac{1}{2}} L^{\frac{1}{2}} T^{-1} \right] = \left[K^{\frac{1}{2}} M^{\frac{1}{2}} L^{\frac{3}{2}} T^{-2} \right]$$

$$\therefore \quad \left[\mu^{-\frac{1}{2}} K^{-\frac{1}{2}} \right] = \left[LT^{-1} \right]$$

i.e., $$\frac{1}{\sqrt{\mu K}} = \text{a velocity} = c$$

where c is a constant.

Experimentally it may be shown that :—

$$1 \text{ E.M.U. of current} = 3 \times 10^{10} \text{ E.S.U's of current} \qquad \text{(3)}$$

Let i_e be the number of E.S.U. of current in a given current,

i_m be the number of E.M.U. of current in the same current.

$$\text{Then} \quad i_e \left[K^{\frac{1}{2}} M^{\frac{1}{2}} L^{\frac{3}{2}} T^{-2} \right] = i_m \left[\mu^{-\frac{1}{2}} M^{\frac{1}{2}} L^{\frac{1}{2}} T^{-1} \right]$$

$$\therefore \quad \frac{i_e}{i_m} \left[LT^{-1} \right] = \left[\mu^{-\frac{1}{2}} K^{-\frac{1}{2}} \right] \qquad \text{(4)}$$

$$\text{But from (3)} \quad \frac{i_e}{i_m} = 3 \times 10^{10}$$

$$\therefore \text{ from (4)} \quad \frac{1}{\sqrt{\mu K}} = 3 \times 10^{10} \frac{\text{cms}}{\text{sec.}}$$

= the velocity of light.

4 **Near and Distant Fields.**—Fig. 3 represents vectorially the direction of the associated fields at points near the transmitter and at a distance.

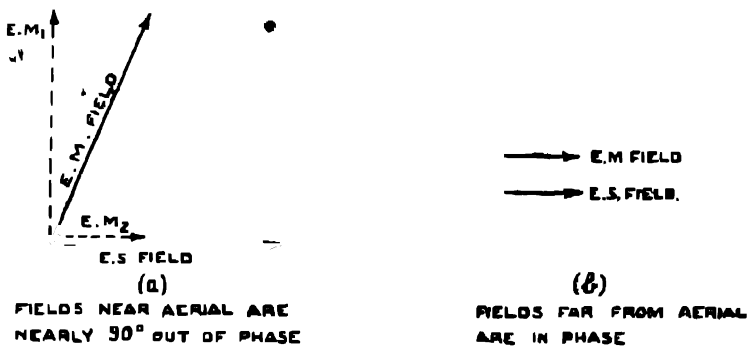


FIG. 3.

In Fig. 3 (*a*) with reference to the electrostatic field, E.M.$_2$ represents the time phase of the radiative component of the magnetic field, and E.M.$_1$ represents that of the inductive component. The time phase of the complete electro-magnetic field in the neighbourhood of the aerial is therefore the resultant of these two fields. It is shown below, that at distances exceeding about 5 wavelengths, the intensity of the inductive field E.M.$_1$ is only about 1/30 of that of the radiative fields and so may be neglected in comparison with the latter. Fig. 3 (*b*) represents the state of affairs at some distance from the transmitter, and shows the two fields in time phase. The intensity of these fields is further referred to quantitatively in paragraph 7.

5. **Polarisation, Wave Front, Wavelength.**—Considering the detached loops of Fig. 1 (*d*), all points along the surface HQR experience the same moving flux at the same time. HQR represents the " **wave front,**" which is the surface joining together all points that experience the same flux at the same time. The direction of propagation is always at right angles to the wave front, though it may go forward or backward, a matter which depends upon the relative direction of the electric and magnetic fields.

The velocity with which the whole system moves outward is approximately 3×10^8 metres/sec. The detached loops move outwards with ever-increasing height but preserving a constant width. The frequency with which consecutive loops are generated is the frequency of the current in the

oscillatory system, and the radial distance between two consecutive maxima of electric or magnetic field in the same direction is the " **wavelength** " of the electro-magnetic wave. In Fig. 1 (*d*) points O and Q are one wavelength (λ) apart. It can be shown that with a sinoidal aerial current the intensity of the electric or magnetic field also varies sinoidally with respect to time at any given point remote from the transmitter. With constant velocity of propagation this implies that the intensity of the fields will vary sinoidally with respect to distance measured outwards from the transmitter at any given time. A sine wave graph, with distance from the transmitter along the " X " axis, can then represent the instantaneous intensities of the fields at different distances, and provides a reason for the use of the term " **wave motion.**" The attenuation in amplitude which takes place during propagation is here neglected, but will be referred to later.

It follows that for a stationary observer situated at H, the frequency with which successive maxima will follow each other is given by the relation

$$f = \frac{c}{\lambda} \quad \text{or,} \quad c = f\lambda,$$

where c is the velocity of propagation, λ is the wavelength, and f is the frequency.

Where the electrostatic flux lines are vertical, as in the direction OQ of Fig. 1 (*d*), the wave is said to be vertically " **polarised.**" When the electric field is horizontal, the wave is said to be horizontally polarised. In these cases vibration of the " æther particle " undergoing the wave motion may be said to be either up and down or backwards and forwards, and so it is in linear vibration. The plane which contains the electric vector and the direction of propagation is called the " **plane of polarisation.**" Actually this is a conventional matter, and some writers understand by this term, the plane which contains the magnetic vector and the direction of propagation. In certain circumstances, however, more complicated motion of the æther is produced. For instance, the æther particle may move in a circle or in an ellipse. The wave is then said to be circularly or elliptically polarised. These various types of polarisation can be illustrated by fixing a long rope at one end. If the free end is then waved up and down, the wave motion in the rope is vertically polarised; if backwards and forwards horizontally, a horizontally polarised wave is produced. Circular polarisation may be imitated by moving the free end of the rope in a vertical circle. All electro-magnetic waves are " **transverse waves,**" because the electric and magnetic vectors are at right angles to the direction of propagation. This distinguishes E.M. waves from sound waves, which are projected through air or some other medium, and are " longitudinal " in nature.

6. **Numerical Example.**—It may help to fix ideas of the nature of the wave and its propagation if a numerical example is taken. Consider a wave of frequency 250 kc./s., and therefore of wavelength 1,200 metres in the free æther, and approximately of this wavelength when travelling through the atmosphere close to the surface of the earth. The electric and magnetic fields are in time phase, and at some instant they will have their maximum values simultaneously at the point O in Fig. 2. Suppose that times are reckoned from this instant. The time of a complete cycle of values of the wave fields for a 250 kc./s. wave is $\frac{1}{250{,}000}$ second, or 4 microseconds. After a quarter of this time, *i.e.*, one microsecond, the two fields at O will have decreased to zero. They then reverse in direction, and start to increase in magnitude, and after another millionth of a second they attain maximum values in the reverse direction. In this time, 2 microseconds, the wave has travelled 3×10^8 metres per second $\times 2 \times 10^{-6}$ seconds $= 600$ metres, or half a wavelength, and so the fields at a point P, 600 metres from O in the direction of propagation, have their maximum values 2 microseconds after similar maxima occur at O. As this time corresponds to maximum reversed values at O, it follows that the fields at P and O at the same instant are 180° out of phase.

In 4 microseconds the fields complete a cycle, and their values at O are the same as their initial values. The wave has travelled 1,200 metres in this time to a point Q. The fields at O and Q, 1,200 metres apart, are therefore in time phase. [*Cf.* points O and Q in Fig. 1 (*d*).]

Thus the time lag of the fields at a point R on those at O depends on the time taken by the wave to travel from O to R. 1,200 metres from O, the fields are in phase with those at O: 600

metres from O, they are 180° out of phase and 300 metres from O they are 90° out of phase. In the general case, the time lag of the fields at R on those at O is $360d/\lambda$ degrees, or $2\pi d/\lambda$ radians, where λ is the wavelength and d is the distance between O and R.

7. **Quantitative Relations.**—Fig. 4 represents a small portion of a radiating wire carrying a current given by $\mathcal{J} \sin \omega t$.

By considering the effect on the form of the electric flux lines when a charge is accelerated at the radiator, and on the assumption that any effect at P will be observed at a finite interval of time after some cause at O, the total magnetic field at a distant point P may be determined, and is expressed by—

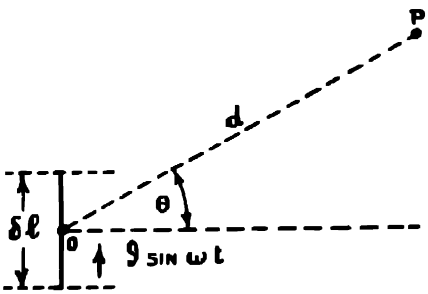


FIG. 4.

$$h = \frac{2\pi f.\ \delta l.}{dc} \cdot \mathcal{J} \cos \omega \left(t - \frac{d}{c} \right) \cos \theta$$

Radiative term

$$- \frac{\delta l\ .\ \mathcal{J}}{d^2} \sin \omega \left(t - \frac{d}{c} \right) \cos \theta$$

Inductive term

All quantities are in E.M. or C.G.S. units. The mathematical investigation which produces the above formula, merely gives a precise mathematical form to the basic ideas of radiation somewhat hazily sketched in paragraph 1.

Of the two terms in this formula, the mathematical treatment shows that only the first is produced by the acceleration of a charge; the second term simply represents the ordinary field that causes the self-inductance of the aerial, and could have been otherwise determined.

Considering the inductive term we note that :—

(*a*) Its amplitude does not depend upon f, the frequency of the current in the radiator.

(*b*) It is in time phase with the current that causes it, after making due allowance (d/c) for the time required in propagation.

(*c*) The amplitude varies inversely as the square of the distance from the radiator and accordingly is only of importance in the immediate neighbourhood of the aerial. Its amplitude will equal that of the radiative field numerically, at a distance given by

$$\frac{2\pi\ .\ \delta l}{d\lambda} = \frac{\delta l}{d^2}$$

$$d = \frac{\lambda}{2\pi} \doteqdot \frac{\lambda}{6}$$

From this it follows that at a distance of 5λ, the intensity of the inductive field is only about 1/30 of that of the radiative field, and may be neglected in comparison with it.

(*d*) When the frequency of the alternating current becomes zero, that is to say, when the current becomes an unvarying one, the formula for the inductive field in a direction at right angles to the radiator reduces to the expression $\frac{\delta l\ .\ \mathcal{J}}{d^2}$.

Considering the radiative term we note that :—

(*a*) It is 90° out of time phase with the inductive magnetic field, and with the current that causes it, if due allowance (d/c) is made for the time required in propagation. It must, therefore, be in time phase with the radiative electrostatic field.

(*b*) The amplitude varies inversely as the first power of the distance and accordingly does not fall off so rapidly as the inductive field.

(*c*) The amplitude is directly proportional to the frequency f, the height of the radiator δl, and to I, the R.M.S. value of the aerial current assumed uniform over the small length of the aerial considered in Fig. 4. The form of this term therefore provides the reason why the frequencies in use for radio communications work must be high, and for the use of high open aerial systems with large uniform currents in the vertical radiating part. No appreciable radiation of energy takes place until the frequencies exceed about 10,000 cycles per second.

(*d*) Associated with the radiative magnetic field there is the equivalent electrostatic field, in time phase with it, as suggested in paragraph 1. From the relation $\mathscr{X} = c\mathscr{H}$, the expression for x becomes

$$x = \frac{2\pi f \, . \, \delta l}{d} \, . \, \mathscr{I} \cos \omega \left(t - \frac{d}{c} \right) \cos \theta \quad .. \quad \text{in E.M. units.}$$

There is also a radial electrostatic field which dies out rapidly like that of the inductive magnetic field.

8. **Total Power Radiated.**—With reference to Fig. 5 the radiation from a dipole may be summed up by assuming it to be placed at the centre of a *large* sphere, the surface of which continually will be penetrated by the outgoing moving fields. In elementary magnetism and electricity it was shown that electric and magnetic fields really represented the storage of energy, numerically evaluated from the well-known formulae $\frac{1}{2}CV^2$ and $\frac{1}{2}LI^2$ respectively. Both of these may easily be put into forms more readily useful, and reduce to—

$$\text{Energy in the electric field} = \frac{KX^2}{8\pi} \quad .... \quad \text{per unit volume.}$$

$$\text{Energy in the magnetic field} = \frac{\mu H^2}{8\pi} \quad .... \quad \text{per unit volume.}$$

All symbols have their usual meanings. Since these two energy densities are equal and complementary, the total energy density may be written

$$\frac{KX^2}{8\pi}(1 + 1) \; .... \; \text{per unit volume.}$$

Having arrived at an expression for X, the electric field at such a distance from the transmitter that the inductive field may be neglected, we can add up mathematically the energy flowing through the whole surface of the sphere. Only the result of this " integration " will be given here.

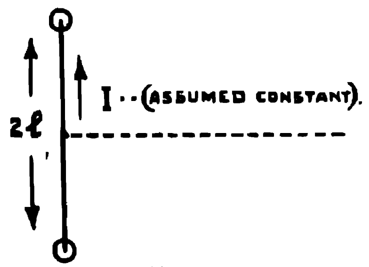


FIG. 5.

Total power radiated $= \dfrac{320\pi^2 l^2 I^2}{\lambda^2}$; the result will be in watts, if l and λ are in the same units and I is in amps. For reasons to be explained more fully later, it is convenient to use the symbol l to denote one-half of the length of the dipole—Fig. 5. It is also noted that " all-round " radiation is provided in the horizontal plane, and that the effect of the " cos θ " part of the radiative term of paragraph 7 is that there is no vertical radiation. The horizontal and vertical polar diagrams of field strength are shown in Fig. 6.

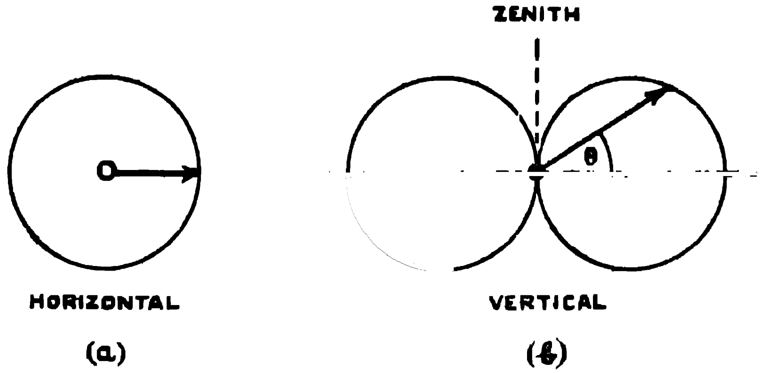


FIG. 6.

9. **Radiation Resistance.**—Since the power radiated is proportional to I^2, if the other quantities remain constant, it may be considered that the power is expended in heating a fictitious resistance R_r, such that—

$$I^2 R_r = I^2 \left(\frac{320\pi^2 l^2}{\lambda^2} \right)$$

whence $$R_r = \frac{320\pi^2 l^2}{\lambda^2}$$

It may be defined as that fictitious resistance which, when multiplied by the square of the aerial current, measures the power radiated.

10. **Radiation Height—Radiation Constant.**—The formula in paragraph 8 for the power radiated, is based on the assumption that the amplitude of the oscillatory current is the same at all points in the radiating wire. This is far from being the case in practice, and it will be shown later that, considering only one-half of the dipole, the current distribution from the centre to the upper end may be either approximately linear, as in Fig. 7 (*a*), or sinoidal as in Fig. 7 (*c*).

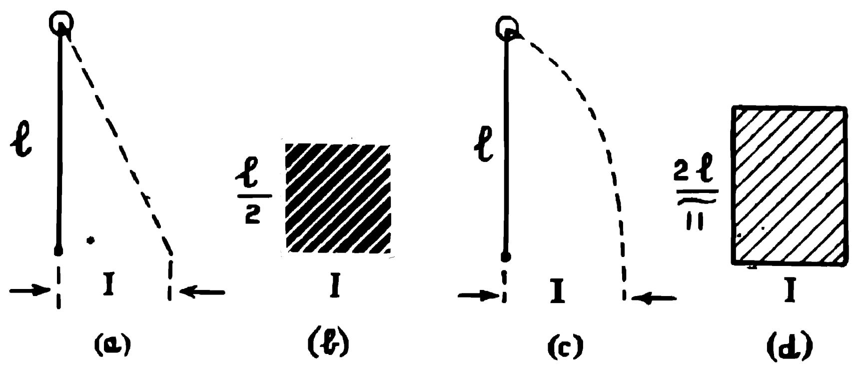


FIG. 7.

The effect is to reduce the amount of the power radiated, and to make the aerial equivalent to a shorter one with the same current. The actual length of the equivalent shorter aerial may easily be assessed by finding the " mean value " of I over the length concerned.

In Fig. 7 (*a*), the mean value is I/2, and the equivalent height is hence $l/2$. In Fig. 7 (*c*), the mean value of I sin θ from 0° to 90° may be shown mathematically to be $2I/\pi$. From this it follows that the equivalent height in this case must be $2l/\pi$. These "equivalent heights" are often called "RADIATION HEIGHTS." Figs. 7 (*b*) and 7 (*d*) represent the state of affairs in ideal equivalent aerials having a uniform current in them equal to the maximum current. They also make clear the importance of the product "lI" in transmitting aerials; diagrammatically, it is the area of a rectangle. It is quite common practice to indicate the power of a station in "METRE AMPERES," *i.e.*, the radiation height multiplied by the maximum R.M.S. current. The expression l_rI, where l_r is the equivalent or radiation height, is termed the "RADIATION CONSTANT." It is clearly a more accurate method of description than that by which a station is referred to as being of so many kilowatts. The latter expression leaves doubt as to whether the power input to the generator, to the wireless plant, or to the aerial is meant.

The formula of paragraph 8 for the total power radiated by a simple dipole will therefore lead to high results unless l_r is used.

Using l_r, we have for the power radiated by a dipole

$$P = \frac{320\pi^2 l_r^2 I^2}{\lambda^2} \quad \text{and} \quad R_r = \frac{320\pi^2 l_r^2}{\lambda^2}.$$

When l_r cannot be evaluated approximately in the way shown above, it can be estimated with the help of a "form factor" which depends upon the design of the aerial system, or it can be measured by determining the field strength at a point in the neighbourhood of the transmitter. Both of these matters are discussed in later paragraphs.

11. Radiation Resistance in Special Cases.

(*a*) THE HERTZ $\lambda/2$ AERIAL.—It is clear that the physical height of an aerial may be expressed in terms of the wavelength in use. In the case when the whole dipole is a half wavelength, putting $\lambda = 4l$ ...... and taking l_r as $\frac{2l}{\pi}$, then from

$$R_r = \frac{320\pi^2 l_r^2}{\lambda^2}, \text{ we have } R_r = \frac{320\pi^2\, 4l^2}{16l^2\pi^2} = 80 \text{ ohms.}$$

This value is roughly constant for all "$\lambda/2$" aerials remote from the earth, and provides a simple rule which enables radiated power easily to be estimated in this very important case.

(*b*) THE MARCONI $\lambda/4$ AERIAL.—It will shortly be shown that, in many cases, the radiating aerial is equivalent to only the upper half of the simple dipole of paragraph 8. The whole aerial is then only a quarter wavelength, and in that case the power radiated is given by

$$P = \frac{160\pi^2 l_r^2 I^2}{\lambda^2} \quad \text{and} \quad R_r = \frac{160\pi^2 l_r^2}{\lambda^2} \doteqdot \frac{1580\, l_r^2}{\lambda^2}$$

(a well-known formula).

Again putting $\lambda = 4l$, and $l_r = \frac{2l}{\pi}$ we find that

$$R_r = 40 \text{ ohms.}$$

This is roughly constant for all $\lambda/4$ aerials, and is another important simple rule which enables radiated power to be estimated.

(c) **AERIAL LENGTH SMALL IN COMPARISON WITH λ.**—Low frequencies are necessary for transmission over long distances, when silent zones must be avoided. At these frequencies it is seldom possible for the aerial height to be at all commensurate with the wavelength in use, with the result that the radiation resistance is invariably very low. There is obviously a practical limit to the height of an aerial. If this is taken to be about 250 metres, then for waves of frequencies less than 30 kc./s. (over 10,000 metres), the radiation resistance cannot be greater than $1{,}580 \times \frac{250^2}{10{,}000^2}$, or 1·0 ohm (using the formula given above). At 15 kc./s. (20,000 metres) the radiation resistance will be less than 0·25 ohm, and in one transmitter working on 16 kc./s. the radiation resistance is as low as 0·055 ohm. In that case exceptional efforts were made to reduce the aerial "loss resistance" which was brought down to the value 0·5 ohm. Even so, the aerial efficiency is low and is given by

$$\text{Aerial efficiency} = \frac{\text{Power radiated}}{\text{Power supplied}} = \frac{0{\cdot}055}{0{\cdot}5} = 11\ \%.$$

12. **Communication Range.**—The amount of energy radiated by a transmitter is no criterion as to the distance over which communication is possible. Before the possibility of long distance communication by high frequency waves was realised, attention was mainly confined to the amount of energy radiated from an aerial on different horizontal bearings. The radiation from a vertical aerial, as would be expected, is symmetrical in this respect and takes place equally in all directions, but many aerial systems have been found to give the maximum amount of radiation on one particular bearing, and little or none on others. This is **directional transmission,** and will be discussed later. Furthermore, the increasing use of the indirect ray, the radiation which travels up into the ionosphere in its passage between transmitter and receiver, leads to the necessity of discovering how radiation from an aerial varies at different vertical angles. Successful point-to-point long-distance transmission at high frequency, depends upon a proper choice of the frequency together with the vertical angle at which the radiation leaves the transmitter, and at which it is directed upwards towards the ionosphere.

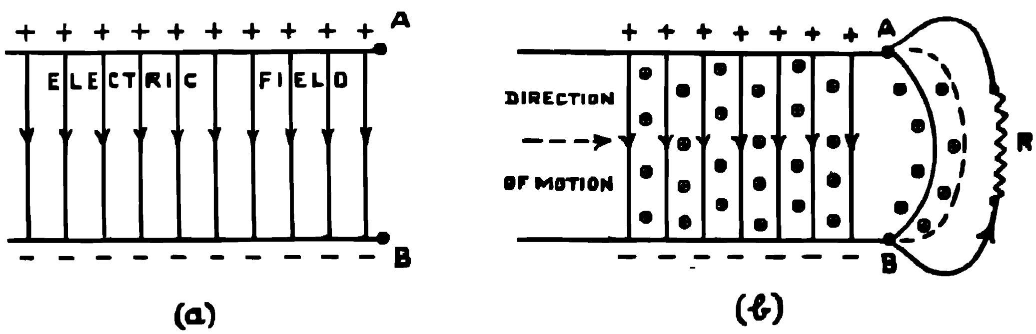


FIG. 8.

13. **Guided Waves in Feeders and Aerials.**—It has been seen that a good aerial involves the simple requirement that a large current should flow between two points as far apart as may be possible. In paragraph 10 it was suggested that, in practice, the amplitude of oscillatory current at different points was unequal, and it is now necessary to examine this matter in some detail.

Maxwell helps us to have a picture of what happens when a current flows in a circuit. He imagined a very large parallel plate condenser, having the two plates oppositely charged to the same potential—Fig. 8 (*a*).

If now a wire of resistance R is joined across AB as in Fig. 8 (*b*), electrons will flow from B to A, or in conventional terms a current will flow from A to B. The collapse of the electric field may be regarded as a drift of the tubes from left to right, the movement being associated with a magnetic field at right angles, due to the " displacement current." The electric flux lines making successive contact with the wire, maintain the P.D. that causes electrons to flow. *A current* of electricity in a conductor, is thus *the result of a movement of the associated electric and magnetic fields*: it is a "*guided wave.*" In a purely resistive circuit the current and voltage are in phase, which implies that the two fields are in phase. The state of affairs is thus similar to that of radiated energy in free space : the velocity with which the tubes collapse from left to right will be that of light. Conversely, the same argument applies to the propagation of electric flux along a pair of infinitely long feeder lines joined to an alternator: if there is no expenditure of energy in the conductors the velocity of the "guided" wave will be that of a free wave—namely 3×10^{10} cms./sec., and Fig. 9 gives a pictorial representation of the process.

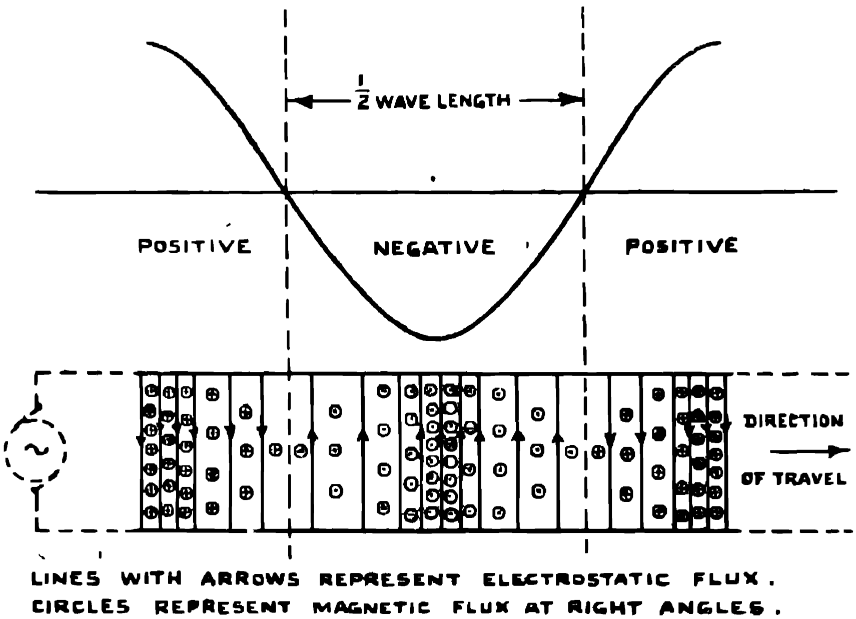


FIG. 9.

At this point it can be emphasised, that the supply of H/F energy to an aerial involves processes essentially similar in nature to those that take place between a power station supplying A.C. at 50 cycles to a distant house. The difference is one of degree, in that in the former case the physical length of the aerial may be commensurate in size with the wavelength of the frequency in use ; this would hardly be possible in the latter case, for the value of λ at 50 cycles is about 3,726 miles, and the power station and the consumer are electrically side by side.

14. **Standing Waves.**—The phenomenon of standing or stationary waves occurs in many branches of science ; it is found in dealing with light waves, with sound waves, and with transverse waves propagated through fluids. The mechanism of stationary waves may be simply understood by considering their production in a rope, one end of which is fixed. If the rope is extended horizontally and the other end is shaken up and down, a **travelling wave** will move along the rope. At the other end, the wave is reflected and travels back to the hand and a system of **stationary waves** on the rope is set up if the agitation is continued. The incident and reflected travelling

waves reinforce each other at certain points and nearly balance each other at other points. The points where practically no up-and-down motion of the rope takes place are called "nodes" and the points of maximum displacement are called "anti-nodes."

The explanation may be put into more definite form by considering the ideal case of a transverse wave motion being propagated through a medium without any loss of energy. Fig. 10 (*a*) shows a wave travelling towards the point A, and Fig. 10 (*b*) shows a reflected travelling wave proceeding in the same medium with the same velocity in the opposite direction. The reflected wave is shown to be 180° out of phase with the incident wave. Fig. 10 (*a*) and Fig. 10 (*b*) represent the same instant of time, and since the resultant displacement of the medium must be the sum of the displacements due to each wave separately, the resultant displacement in this case must be zero everywhere. In spite of this, certain points in the direction of propagation have considerable

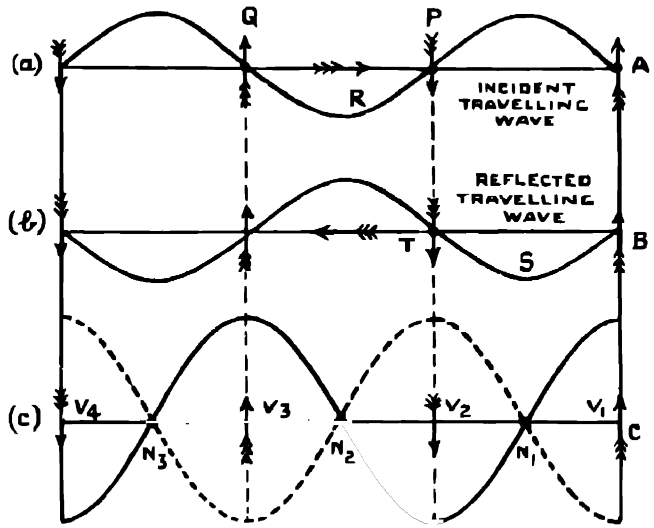


FIG. 10.

transverse velocity. Considering the corresponding points P and T it is clear that the trough R will at some later time arrive at P, and, at the same time, the trough S will arrive at the point T. Accordingly, at the instant of time represented by Fig. 10 (*a*) and (*b*), the corresponding points P and T must both be moving downwards with some velocity V_2. Fig 10 (*c*) represents the state of affairs a quarter of a period later, when the trough S and the trough R are both reinforcing each other, the total displacement being the sum of the separate displacements due to the incident and reflected waves. By considering the state of affairs at subsequent intervals of time, it should be clear that nodal points will always occur at N_1, N_2, N_3. **Anti-nodes** therefore appear as points of maximum displacement and of maximum velocity, whereas **nodes** appear as points of zero displacement and zero velocity.

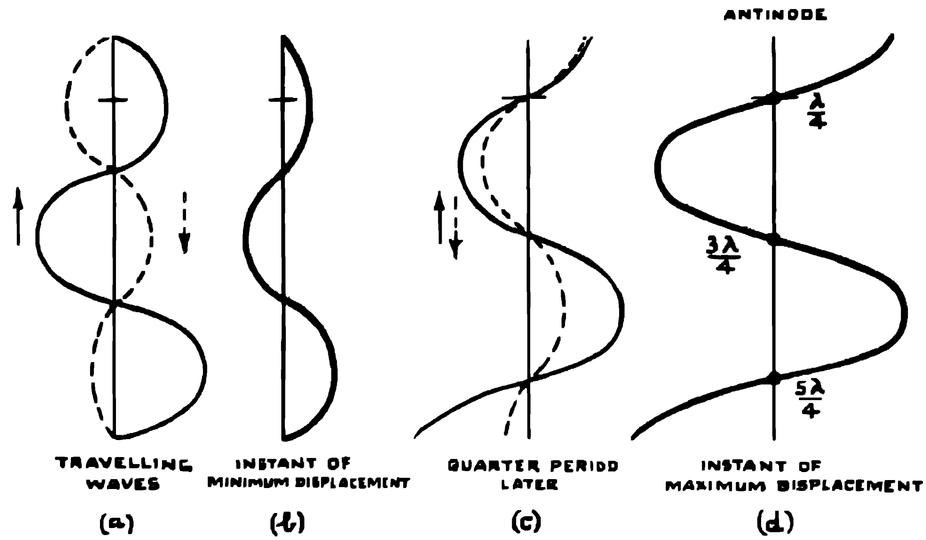
In short, standing waves arise when travelling energy is reflected at a point, as it always must be if it cannot be used: it cannot just disappear.

15. **Standing Waves in a Long Vertical Aerial.**—Fig. 11 represents the upper half of a vertical dipole aerial, of the same nature as Fig. 1 (*a*) or Fig. 4. AB is long in comparison with λ. BC is the coupling coil, and is the source of the travelling waves of voltage and current in the two halves of the aerial. Now the aerial radiates energy and has resistance losses throughout its length, so that the power input to any section DA will be less the nearer D is to A. The peak values of

voltage and current will therefore also decrease, and, finally, reflection of the surplus energy takes place at A.

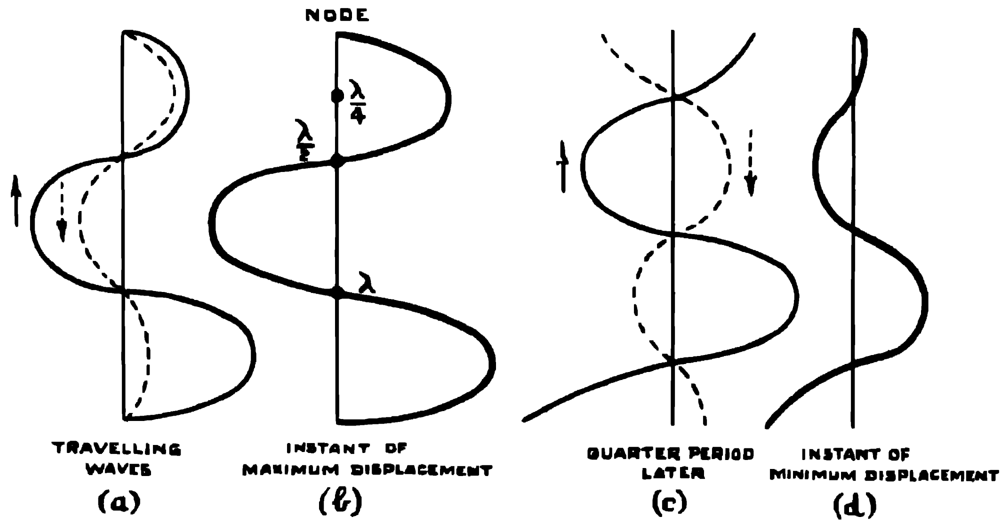
In Fig. 12 (*a*) the full line curve represents the WAVES OF VOLTAGE travelling up the aerial and decreasing in amplitude as they go. When they reach the top they are reflected downwards, and the phase is changed by 180°. This phase change represents the charging of the aerial capacity in the opposite sense. The dotted curve represents the reflected voltage which decreases in amplitude as the wave travels to the bottom. The total voltage to earth at any point is the sum of the upwards and downwards travelling waves for the instant of time chosen. Imperfect stationary

FIG. 11.



VOLTAGE WAVES

FIG. 12.



CURRENT WAVES

FIG. 13.

waves are therefore set up. With perfect reflection and no radiation of energy, at the instant of time of Fig. 12 (*a*) there would be no nett displacement anywhere. Fig. 12 (*b*) represents the actual state of affairs. Fig. 12 (*c*) represents the position a quarter of a period later, the nett result at that instant being shown in Fig. 12 (*d*). There is now a clearly defined anti-node at the top of the aerial, with momentarily nodal points at distances $\lambda/4$, $3\lambda/4$, $5\lambda/4$, etc., down the aerial.

The action of the current waves is somewhat different. In Fig. 13 (*a*), the full line represents the WAVES OF CURRENT travelling upwards and decreasing in amplitude as they go. When they reach the top of the aerial they too are reflected back, but their phase is not changed by 180° as in the case of the voltage waves. This should be clear after reference to Fig. 2, for reversal of any two of the three vectors does not involve reversal of the third. The dotted curve therefore shows the reflected waves travelling downwards. As in the case of the voltage waves, the current at any point at any instant is given by the sum of the two waves. Fig 13 (*b*) represents the nett displacement at the instant of time of Fig. 13 (*a*). It shows a node of current at the top of the aerial and at similar points an even number of quarter wavelengths down the aerial. Anti-nodes of current occur at points an odd number of quarter wavelengths down the aerial. Fig. 13 (*c*) shows the position a quarter of a period later, and, as for Fig. 12 (*a*) and (*b*), it represents the instant at which there would be no nett displacement anywhere if no radiation, etc., occurred: Fig. 13 (*d*) shows the actual state of affairs, and again demonstrates that the nodal points are imperfect ones.

The distance between two points of maximum current or voltage will measure a half wavelength, provided that the velocity with which the waves travel along the aerial is equal to that of light. This is the case if the aerial is a thin wire, but when the aerial is a tube the velocity is decreased, and the distance between two peaks of current will measure less than half a wavelength.

When the aerial becomes very long compared with the wavelength (as is sometimes the case when the main aerial in a ship is used for H/F transmission), the decrease in amplitude of the reflected waves becomes so large that their effect on the travelling waves proceeding from BC may be neglected. The current and voltage waves then correspond to those in the upward travelling wave only, and are therefore in phase In other words, the aerial behaves as a resistance, no matter what the frequency is, provided that it is high enough for the above condition to be fulfilled. Such an aerial would be independent of the tuning and could be used to transmit any frequency. Furthermore, since there are no nodes or anti-nodes at the source, the centre of BC could be *earthed*.

16. **Standing Waves in a Short Vertical Aerial.**—When the whole length of the dipole is of the same order of magnitude as the wavelength itself, the form of the standing waves leads to important results. These are summarised in Fig. 14, the curves being easily derived from Figs. 12 and 13.

THE $\lambda/2$ AERIAL.—Fig. 14 (*a*) shows that there is a voltage node in the centre of a system which is $\lambda/2$ in length; it also makes clear that, with reference to the R.M.S. aerial current, the R.M.S. voltages to earth at the upper and the lower ends are, respectively, 180° out of phase [*cf.* Fig. 1 (*a*)]. It follows, therefore, that the R.M.S. voltage in the neighbourhood of the imperfect node is in phase with the aerial current. These phase relationships are shown vectorially in Fig. 14 (*a*).

The impedance of the aerial is measured by the ratio V/I, at any point, and this is clearly a minimum at the centre. In paragraph 11 it was shown that this impedance is of the order of 80 ohms. Fed in this way, the system is therefore equivalent to a series resonant or acceptor circuit, and is sometimes described as being "current fed." Since it is electrically balanced about the centre, it is, mechanically, something like a see-saw. If, however, the same circuit were fed from a coupling circuit at the top or the bottom, the impedance would be very big, of the order of several thousand ohms, and the whole would behave as a parallel resonant or rejector circuit. Such a system is sometimes described as "voltage fed." It is important to notice that the impedance will vary from a high value at the free ends to a low value at the centre. A wire having on it a standing current wave of this nature, is said to be excited at its "FUNDAMENTAL OR NATURAL FREQUENCY." The phrase is a borrowed one, for the wire is similar to a violin string vibrating as a whole, or to an organ pipe open at the top.

THE $\lambda/4$ AERIAL.—Since any voltage node is "earthy" in potential, it may in fact be earthed

without affecting the voltage and current distribution. This idea led Marconi to the use of a λ/4 aerial, Fig. 14 (*b*). With the help of a coupling coil, the aerial can be excited at the earthed end, and will have a fundamental resonance when the frequency corresponds to a wavelength roughly four times the length of the aerial. As a matter of interest, this may be compared with a closed organ pipe, which when blown produces a note of wavelength approximately four times the length of the pipe. It will be noted that the anti-node of current is at the earthed end, which provides the reason for fitting aerial ammeters at that point for λ/4 aerials.

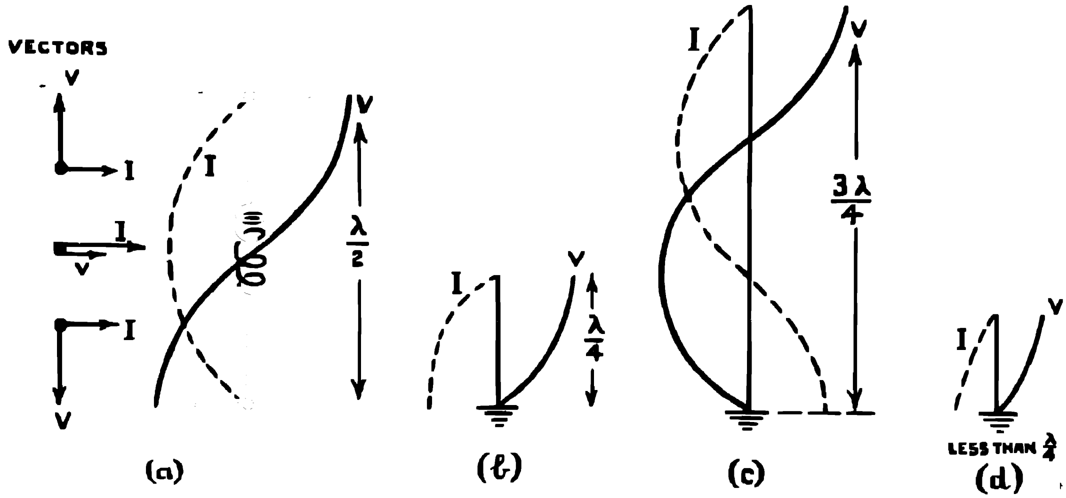


FIG. 14.

The working of a λ/4 aerial may also be appreciated by considering a large horizontal copper sheet POP_1 (effectively the earth) to be inserted at the symmetrical point perpendicular to the Hertzian radiators of paragraphs 1 and 8.

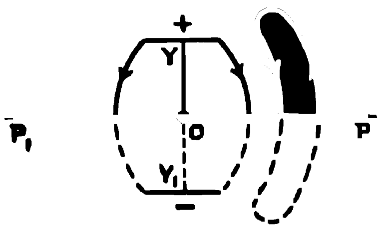


FIG. 15.

Fig. 15 shows lines of electric force near such a radiator. The lines meet the conductor POP_1 at right angles to the surface, and the lower half of the system is electrostatically screened from the upper half. Under these conditions, the insertion of the copper sheet produces no change in the electrostatic field.

If the lower half OY_1 of the radiator is removed, no change is noticeable above POP_1. If POP_1 is a perfect conductor, the currents flowing in it consume no energy In the practical case where the earth is used instead of a perfect conductor, these currents absorb energy from the passing wave front, the lower end of it moves more slowly in consequence, and a forward tilt results. Fig. 16 shows diagrammatically the radiated fields, together with the associated earth currents produced by the radiation from an earthed λ/4 aerial. The forward tilt assists L/F waves to follow the curvature of the earth (paragraph 28). The earth currents will be in the form of circular bands, within which the current will flow, alternately, radially outwards and inwards.

A comparison may now be made between the action of a λ/2 dipole in FREE SPACE and an earthed λ/4 aerial, each working at the same frequency and having the same maximum R.M.S. aerial current. In both cases, there will be the same field strength at corresponding points P; the λ/2 aerial will, however, radiate twice as much energy—equal amounts above and below the non-existent POP_1 conducting plane.

The invention of the λ/4 earthed aerial, where the earth is one plate of the condenser, is considered to be the most important of Marconi's early contributions to radio engineering. From the short Hertzian radiator he produced lofty and efficient aerial systems, and with the ground ray

working with low radio frequencies, he achieved communication over long distances. Chief among the advantages of this aerial is the fact that it can easily be adapted to cover a range of frequencies.

Hertz and Lodge, both early experimenters on high frequencies, used dipoles. It is interesting to observe that with the increasing use of high frequencies, the $\lambda/2$ dipole is again being much used, and is perhaps the most popular of the H/F aerials.

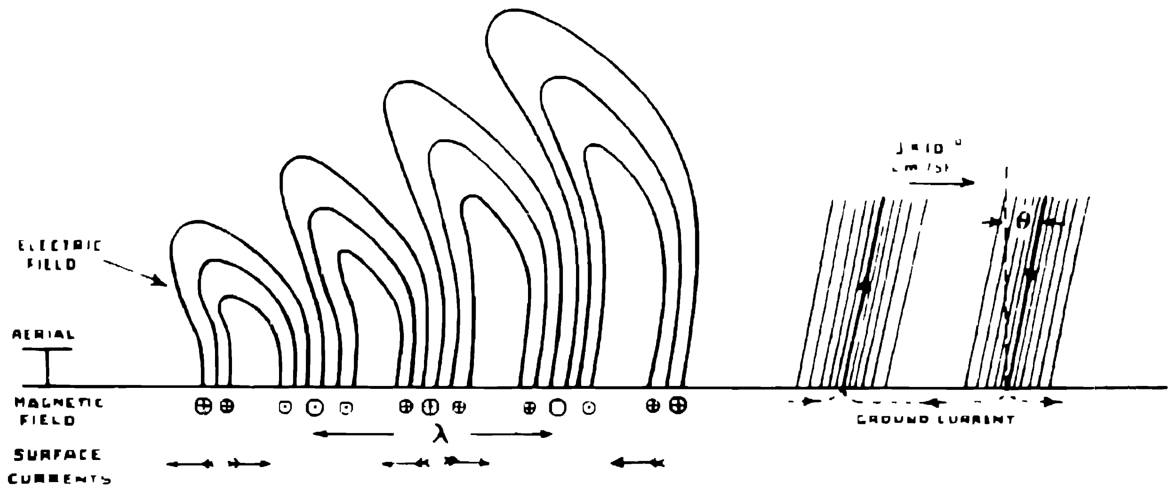


FIG. 16.

There are thus two fundamental types of simple aerial :—

$$\text{THE HERTZ OR } \frac{\lambda}{2} \text{ AERIAL.}$$

$$\text{THE MARCONI OR } \frac{\lambda}{4} \text{ AERIAL.}$$

All other types of simple aerial may be regarded as derivatives of one or other of these.

17. Image Aerials—The Function of the Earth.—The lower half OY_1 of the dipole of Fig. 15 may be regarded as being both the optical and the electrical " image " of the upper half. The conception of an image of an aerial system below the surface of the earth is especially useful in accounting for the effect of reflection from its surface. As in the case of optical reflection from a mirror, the image may be regarded as the centre from which radiation virtually emanates. This is equivalent to saying that the effect of the ground in the neighbourhood of an aerial may be assessed by replacing the ground by an image of the aerial, from which half of the available energy is radiated, and then calculating the radiated field produced jointly by the aerial and its image. The only approximation involved in such a process is the assumption that the earth is a perfect reflector and absorbs no energy.

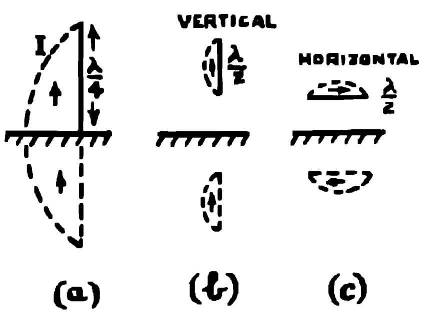


FIG. 17.

Fig. 17 shows an image pattern produced by an earthed $\lambda/4$ aerial, and by an elevated vertical and horizontal $\lambda/2$ dipole.

In cases where it is difficult to obtain a good low resistance earth connection, it is necessary to improve the conductivity of the earth in the neighbourhood of the aerial by burying a vast network of wire. In some cases, where it is not thought possible to make an effective earth, the network of wire is insulated from it and forms the second plate of the condenser of which the aerial wire is the other plate; this form of "earth" is sometimes called a **counterpoise,** or earth screen.

A portable receiver usually needs no earth connection at all, since its aerial system is fundamentally different. It usually employs a loop aerial, the principle of which is fully discussed in the section on "Direction Finding." It should be noted, however, that the phrase "earthed" is frequently used to denote a connection to the case containing the apparatus, and the latter may or may not be joined to the earth. In the case of Service receivers, however, the H.T. negative and L.T. negative are combined, and this common negative is earthed before the leads reach the set.

A portable transmitter, using a $\lambda/4$ or equivalent aerial, may also need no earth connection, though better results will sometimes be obtained if one is used.

18. **Standing Waves.—A Case of Resonance.**—It has been observed that a $\lambda/4$ aerial excited at the earthed end, and a $\lambda/2$ aerial excited in the middle, both behave as series resonant circuits It would therefore appear that all that has so far been explained in terms of standing waves could equally well be accounted for in terms of "line resonance." The voltage anti-node at the top of a $\lambda/4$ aerial may be regarded as similar in origin to the rise of voltage observed across a condenser in a simple resonant circuit. It is, however, not so easy to show in this way that the distributed capacity and inductance of an aerial will make it tune to an applied frequency, when its physical height "l" is related to the wavelength "λ" by the expression

$$\lambda = 4l.$$

More accurately, the factor is between 4·2 and 4·5, in the case of an aerial with perfectly distributed inductance and capacity.

From the mathematical point of view, this method of approach has many advantages.

A most important case of this line resonance is sometimes noted in H/F transmitters of the parallel fed variety. The oscillatory frequency at some part of the range may either be the same as the resonant frequency of the anode choke with it self-capacity, or a harmonic of it, and standing waves may be set up in the choke of sufficient amplitude to damage its insulation. There is also the difficulty that when the choke tunes as a series resonant circuit, the oscillatory circuit will be virtually short circuited; when the choke tunes as a parallel resonant circuit, a voltage anti-node will be formed on the anode of the transmitting valve, which will accordingly be worked with a higher P.D. between the anode and the filament than the rating of the valve will allow.

19. **Aerial Reactance.**—Fig. 18 (*a*) represents an aerial of fixed height attached to a source of supply, the frequency of which may be varied. The curves of Fig. 18 (*b*) represent the variation of reactance with frequency, measured at the earthed end by the ratio V/I.

When the frequency is low, and the length of the aerial is a small fraction of λ [Fig. 14 (*d*)], the aerial acts as a capacity and gives a high value of capacitive reactance OB. As the frequency is increased, the value of the capacitive reactance $\left(\frac{1}{\omega C}\right)$ decreases to zero value when the aerial measures a quarter wavelength. On increasing the frequency still further, the aerial becomes inductive $\left(\omega L > \frac{1}{\omega C}\right)$ and tends towards infinite inductive reactance O_1 A as the aerial approaches a half wavelength. With a slight further increase in frequency, the aerial becomes of infinite capacitive

reactance O_1B_1 and follows through the same cycle, approaching zero for a $3\lambda/4$ aerial, and infinite inductive reactance for a full wavelength aerial. It is clear, that first a quarter, then a half, and finally a whole wavelength has been fitted in turn on to the aerial. The process could be continued. Allowing for losses, the dotted curve shows approximately the variation of effective impedance at the lower end.

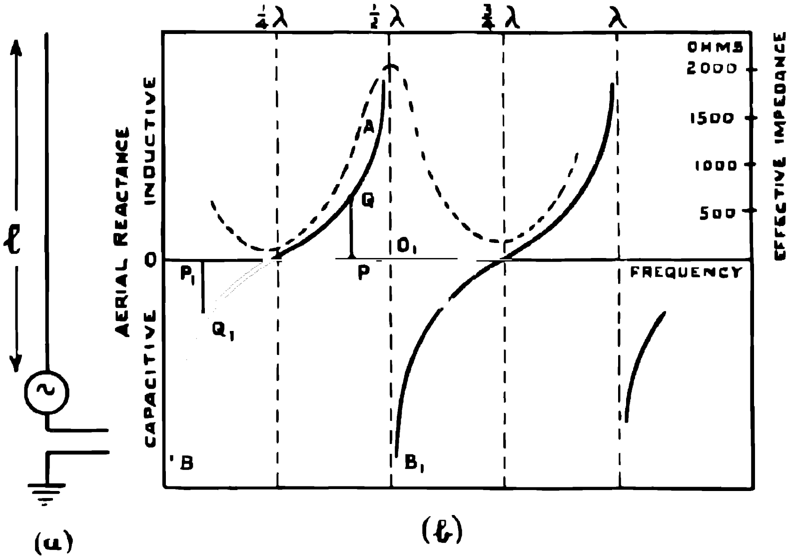


FIG. 18.

20. **Aerial Tuning.**—Assume that the aerial of paragraph 19 is to be tuned to operate at a frequency "f" represented by OP of Fig. 18 (*b*). At this frequency, the aerial inductive reactance is PQ. If the condenser to be inserted in the aerial is of such value that its capacitive reactance is numerically given by PQ, then the aerial reactance will be brought to zero, and the aerial system will be in tune. If, however, the frequency chosen is such that P is very near to O_1, the inductive reactance PQ may be so large that any possible condenser has far too large a minimum value to tune the aerial. If the frequency be represented by OP_1, the aerial reactance $P_1\,Q^1$ is capacitive, and the aerial can only be tuned by inserting an equivalent inductance. In this case, also, the tuning inductance required becomes impossibly large if the frequency is reduced too much. The same applies to the other curves representing the second, third, etc., harmonic modes of oscillation.

The operation of tuning an aerial by adding a " loading " inductance or " shortening " capacity is, therefore, equivalent to adjusting the system to be the *electrical equal* to a $\lambda/4$ aerial.

The main transmitting aerial of a battleship might have a length of 450 ft., giving it a natural frequency of about 500 kc./s. (LC value = 91·2 mic. jars).

For frequencies below 500 kc./s. it could be made into a reasonably efficient $\lambda/4$ aerial, down to about 100 kc./s., by loading it with an inductance. Transmission might in fact be possible down to 60 kc./s. ; voltages at the deck insulator would be very high, brushing would be likely to occur, and aerial efficiency would be very poor.

From the point of view of reception, the above aerial could, of course, be used down to much lower frequencies.

From 500 kc./s., to somewhere near 1,000 kc./s., the aerial could be tuned with the help of a series condenser. Transmission at 1,000 kc./s. could only take place with the help of a suitable high impedance coupling between the base of the aerial and earth. Between some frequency above 1,000 kc./s. to another one somewhere below 2,000 kc./s., the aerial could be operated as a $3\lambda/4$ aerial.

In the case of an earthed $\lambda/4$ aerial that has to be loaded in order that standing current waves may have the largest possible amplitude, the curves of voltage and current distribution will be something like Fig. 19 (*a*) and (*b*).

The loading impedance disturbs the form of the standing waves. At low frequencies the length of the transmitted waves is usually very large in comparison with the length of the aerial, and it has to be tuned with an inductance known as the "aerial tuning inductance." Owing to the concentrated nature of this inductance and its negligible capacity to earth, the potential rises sharply and uniformly across it. For this reason an aerial coil always requires a greater clearance from earth at its upper than at its lower end. For the same reason, the R.M.S. current remains appreciably constant across the coil. Similarly, Fig. 19 (*b*) shows the form of the curves when tuning is effected by adding a capacity in series.

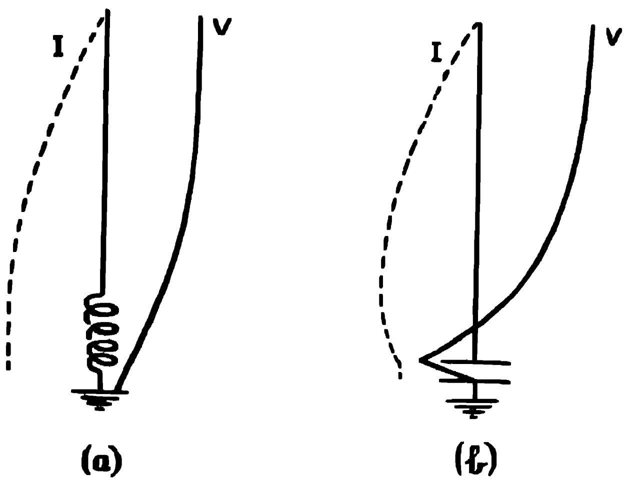


FIG. 19

As has already been seen in paragraph 15, the use of the ship's main aerial for H/F transmission may not involve the production of standing waves extending to the bottom of the aerial. In Service practice afloat, these aerials are always earthed, and when transmitting at H/F the aerial is tuned by means of a condenser and an inductance with a series/parallel switch, until the aerial ammeter shows the maximum current at the earthed end. This is equivalent to adjusting the aerial system to be an odd number of quarter waves in length. If no aerial current is obtained in the "series position," the inference is that the aerial is an even number of quarter waves in length, and, since the voltage anti-nodal point is earthed, no power reaches the aerial. The aerial condenser must be placed in parallel, in order to produce a rejector circuit that may produce the requisite high impedance coupling between the base of the aerial and earth.

The lower frequencies in the H/F range usually require the series position of the series/parallel switch, in order to increase the capacitive reactance and tune the whole system as an acceptor circuit. The higher frequencies in the same range may require either the "series" or "parallel" position, according to whether the untuned aerial is approximately the electrical equivalent of a quarter wavelength or a half wavelength. When an aerial has an electrical length approximately equal to an even number of quarter wavelengths, it can only be energised at its base by means of a high impedance coupling, such as a rejector circuit; including the latter, the lower end of which may be earthed, the whole system then becomes electrically equivalent to an odd number of quarter wavelengths.

In the Service, in all cases where the aerial excitation is indirect, the aerial is tuned until the aerial ammeters show the maximum current. With direct aerial excitation, the aerial tuning is fixed by frequency considerations and the adjustment to maximum aerial current is made by control of the other circuit variables, grid tuning and coupling, etc.

21. **The Roof of an Aerial—Aerial Capacity.**—If a portion of a vertical aerial is bent over horizontally to form an inverted "L," the voltage and current standing waves will be something like Fig. 20. The effect of this addition to the aerial capacity is to increase the mean value of the R.M.S. current flowing in the vertical part of the aerial. This produces an increase in the effective height of the aerial and is therefore much to be desired. A "T" shaped aerial gives a similar result from this point of view, and produces a slightly better broadcast of its radiation. The inverted "L" aerial has very slight directional properties which will be referred to later.

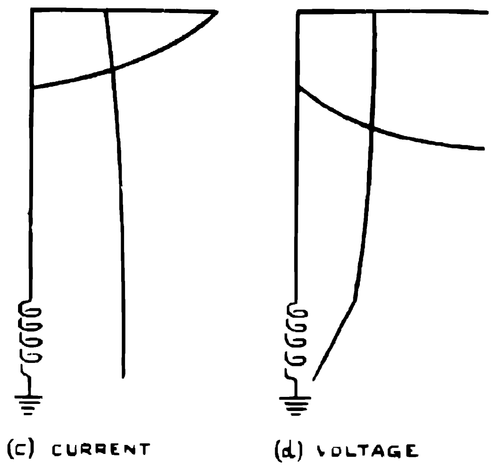


FIG 20.

22 **Aerial Form Factor.**—It is clear that in the case of inverted "L" and "T" shaped aerials, the distribution of current in the vertical part is then no longer approximately sinoidal, and it is impossible simply to arrive at an estimate of the radiation in the way outlined in paragraph 10. It is, however, possible to arrive at a factor by which the actual height in either case may be multiplied in order to obtain the effective height.

Effective height = aerial form factor × actual height.

Let L = length of horizontal part of ┌ aerials.

= half total length for flat-top T aerials.

Let l = height of aerial.

Work out the ratio $\frac{L}{l}$.

Then the A F F may be determined from the following table :—

$\frac{L}{l}$	A.F.F.	$\frac{L}{l}$	A.F.F.
0·0	0·639	1·5	0·940
0·1	0·696	2·0	0·958
0·2	0·741	3·0	0·979
0·3	0·777	4·0	0·987
0·4	0·806	5·0	0·993
0·5	0·830	6·0	0·996
0·6	0·850	7·0	0·998
0·7	0·867	8·0	0·999
0·8	0·881	9·0	0·999
0·9	0·893	10·0	1·0
1·0	0·904	—	—

This factor gives a fair approximation to the radiation or effective height for shore station aerials, and for merchant ships with wooden topmasts of equal height, small metal superstructures and the feeder well away from the funnel.

23. **Aerial Capacity in L/F Aerials.**—An L/F aerial is almost necessarily of the earthed $\lambda/4$ type. With aerials of this nature there are at least three good reasons why the aerial capacity between the roof and the earth should be as large as possible :—

(a) The radiation resistance is proportional to the square of the effective height ; the latter is made large, firstly by making the actual height large, and secondly by attaching to the top of the aerial a horizontal roof which represents a large concentrated capacity, and which, therefore, makes the effective height more nearly equal to the actual height.

(b) It was observed in paragraph 20 that the loading impedance disturbs the form of the standing waves. In a ship, the loading impedance will usually be below deck in the transmitter, and connection to the aerial will be made through a vertical aerial trunk to a deck insulator. If the oscillatory potential of the aerial at this point is much above the potential of the deck insulator, it is possible that a loss due to "brushing" may occur. The greater the capacity of the aerial, the less inductance will be required for tuning to any particular frequency, and consequently the less will be the oscillatory potential of the aerial at the deck insulator. The aerial inductance may be considered to be concentrated in the loading inductance, and the voltage across it is given by

$$E_L = \omega LI \; ; \quad \text{but } \omega = \frac{1}{\sqrt{LC}}$$

$$\therefore \quad E_L = \frac{LI}{\sqrt{LC}} = I\sqrt{\frac{L}{C}}$$

From this it is clear that C should be big in order that E_L may be small.

"Brushing" as distinct from "sparking" is a phenomenon with which every telegraphist is familiar ; it is manifested by a silent bluish discharge from any high potential point in the aerial circuit in the direction of the closest earthed object ; or it may proceed from the earthed point in the direction of the high potential point. It indicates that the air in the neighbourhood of the conductor is being ionised, and its insulation broken down. An aerial brushes most vigorously from its ends and from sharp points and angles If the feeder is immediately abaft the funnels, the smoke and funnel gases will considerably increase its tendency to brush, as they partially ionise the air and decrease its dielectric efficiency. They will also tend to make the aerial wire brittle. For this reason it is best, when possible, to feed the aerial before the funnels, especially in destroyers and other small craft with low aerials. Once an aerial has started to brush, it is useless to increase the aerial current ; the transmitting range will not be increased, and at night the position of the ship will be disclosed.

It is found in practice that this limit is reached, at low frequencies, when the voltage gradient is about 30,000 volts/cm. The potential gradient around an aerial wire depends on its diameter ; the larger the diameter, the less is the danger of brushing. With a single 7/19 aerial wire, the critical voltage gradient might be reached when the oscillatory potential reaches a peak value of about 30,000 volts. For aerials in H M. ships the peak voltage seldom exceeds 40,000 volts.

As an approximation, the loss due to this cause may be taken as inversely proportional to the frequency, for voltages about the critical value.

(c) The capacity of an aerial system is distributed unevenly. In a ship, some of the capacity may be considered to be concentrated between the roof of the aerial and earth, and some between the aerial feeder and the earthed aerial trunk through which it passes from the wireless office.

Fig. 21 shows these two capacities as C_1 and C_2. If C_2 is larger than C_1, the form of the standing current wave will be very much disturbed; most of the aerial current will flow in the wire in the trunk, and little will remain to flow in the elevated radiating wire. The radiation resistance would be seriously reduced.

To retain the condition that a large current should flow between the two points as far apart as may be possible, it is essential that capacity C_1, sometimes called the radiating capacity, should be greater than C_2, sometimes called the non-radiating capacity. The energising of an aerial is often considered to be the operation of charging the distributed capacity to earth of the aerial. In the simple vertical wire, the numerical value of each of the parallel connected capacitive elements decreases as the distance up the aerial increases, quite a small current will charge the capacity element at the top of the aerial to quite a high potential—a voltage anti-node. If C_1 is big, a large current will be required to charge it.

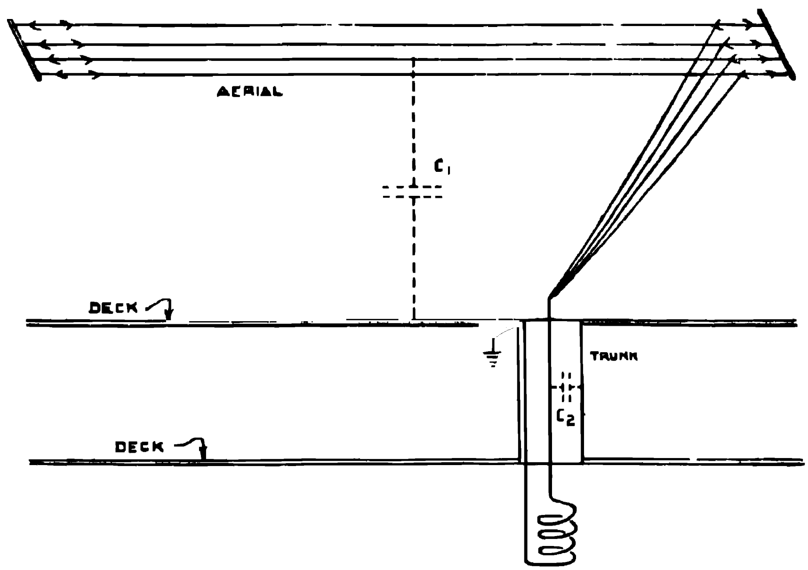


FIG. 21.

In order to have a big radiating capacity, the roof system of wires should have the greatest possible area. Two wires hoisted up parallel to and at a considerable distance from one another, will have a capacity nearly twice that of a single wire, but as the wires approach one another the joint capacity becomes less, so that when they are within (say) a foot of each other, very little extra capacity is gained by the use of the second wire. A few wires spaced far apart are better than many wires near together, so far as total capacity is concerned.

It is bad practice to bring a portion of the aerial low down with the idea of increasing the capacity, although the latter is necessary for frequency range; the decrease in radiation height far outweighs the advantage of any increase in capacity.

The capacity of a ship aerial (σ) will not remain constant unless the aerial is hauled out uniformly at all times. It will vary as the aerial sways in the wind, as turrets are trained, and as awnings are spread or furled. It will also depend upon whether decks are wet or dry, and whether the ship is rolling. The value of σ in a ship varies from 0·3 jar in a submarine to about 2 jars in a big ship. For shore stations, aerials have a capacity up to 10 jars. The value of σ for the big aerial at Rugby is about 40 jars. (The natural *inductance* of a ship's main aerial generally lies between 10 and 70 mics.)

24. **Loss Resistance.**—The effective total resistance of an aerial is made up of two parts—RADIATION RESISTANCE, and LOSS RESISTANCE, the former " useful " and the latter " wasteful."

The loss resistance is due to a number of causes, and it is clearly desirable that this " resistance should be small and the radiation resistance large. The components of loss resistance are summarised in the following table :—

Effective total resistance.

- Loss resistance.
  - Loss in inferior dielectrics.
  - Brushing losses.
  - I^2R losses in aerial and earth circuits.
  - Loss by leakage over insulators.
  - Loss in the ground connection.
  - Loss by eddy currents in neighbouring conductors.
- Radiation resistance.

(*a*) LOSS IN INFERIOR DIELECTRICS.—This loss is due to dielectric hysteresis, an A.C. phenomenon, which is particularly evident in dielectrics such as wood, concrete, masonry, trees, etc., which might happen to be near the aerial and hence acted on by the electrostatic field around it. It is intimately related to dielectric absorption—a D.C. phenomenon.

Smoke or funnel gases do not materially affect the dielectric properties of the air near the aerial ; they may be a contributory cause to the various crackles heard in a neighbouring receiver, but escaping steam is a much more potent one.

(*b*) BRUSHING LOSSES.—These have already been considered.

(*c*) CONDUCTOR LOSSES IN AERIAL AND EARTH CIRCUITS.

THE AERIAL.—The method of construction of the aerial must be a compromise between the requirements for good conductivity and those for mechanical strength.

In order to keep down " skin effect," it is better to use stranded wire than one solid conductor. Bare iron wire should not be used, as, owing to its permeability, " skin effect " is very marked. Galvanised wire is better, as the current will flow over the galvanised skin ; it is not so good as copper wire In most of H.M. Ships the aerial wire is 7/19 phosphor-bronze (not enamelled) ; it is strong but expensive. Insulated wire should not be used, as it becomes covered with a semi-conducting layer of soot which, forming a skin to the wire, would carry a certain proportion of the aerial current and therefore cause a great resistance damping.

The importance of making all joints in the wire with the greatest possible care cannot be too much emphasised. Although a badly-made aerial may send nearly as efficiently as a well-made one, on account of the transmission energy being sufficient to " jump " any small break in the continuity of the conductor, yet, when it is used as a receiving aerial, the minute currents will be unable to flow through any high resistance junction, and great loss of efficiency will ensue. The best arrangement is to make aerial and feeders continuous throughout, the aerial wires being taken round a thimble and used to form the feeders. This arrangement is the strongest and the most efficient electrically.

The construction of an aerial is a troublesome task. Accordingly, when-

ever a new aerial has to be made, it should be done as carefully and strongly as possible, special attention being paid to the measurements of the wire.

The wire should never be soldered in any place which is going to be in a state of tension, since soft solder has little tensile strength. Moreover, soldered joints at thimbles and elsewhere tend to break. This is because the solder makes the stranded wires stiff by filling up the spaces between the strands; continual vibration combined with small flexuring of the wire tends to make it break at the end of the stiff part.

All sharp points, roughnesses, burrs and sharp bends or kinks should be smoothed off or otherwise avoided, because they assist in the leakage of energy in the form of brushing. If these precautions are taken, the aerial, when once up, will remain up for a very long time without giving any trouble.

THE EARTH.—Earth losses depend on the high-frequency resistance of conductors, and the latter increases with the square root of the frequency due to skin effect.

The losses can therefore be assumed to vary directly as the square root of the frequency.

(*d*) LOSS BY LEAKAGE OVER INSULATORS.—Defective insulation will account for a considerable loss in efficiency especially on low frequencies, not only when transmitting, but also when receiving; the loss is greatest during periods of rapid temperature change.

Leakage to earth in this way is equivalent to having a resistance in parallel across the aerial capacity, and this can be replaced by an equivalent series resistance, whose value is given by $\frac{1}{\omega^2C^2R}$, where R is the resistance of the leakage path. Hence losses due to leakage are inversely proportional to the square of the frequency. As an approximation, they are generally taken as varying inversely as the frequency.

The points where losses are liable to occur are at the insulators in the trunk, at the deck insulator, and at outhaul or strain insulators.

It is, therefore, the duty of the wireless staff in a ship to keep all these scrupulously clean, especially after high speed steaming, heavy rain or bad weather, since under these conditions the insulators will probably be covered with a semi-conducting layer of "stokers," dirt, or dried salt.

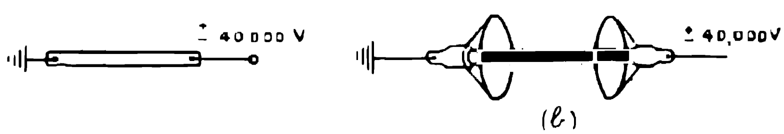


FIG. 22.

Strain insulators are fitted with an "anti-brushing" fitting, as in Fig. 22 (*b*). The reason for this is as follows :—

Suppose we have a plain porcelain rod, as in Fig. 22 (*a*), one end of which is connected to earth and the other end to an aerial wire, which may have an oscillatory potential of 40,000 volts, there will then be a great tendency for sparking to occur over the surface of the rod. If, however, "anti-brushing" rings are fitted, the electric strain is imposed on the air between the two rings, *i.e.*, over a larger area of dielectric, and the tendency to brush is reduced.

The number of strain insulators used in staying out the aerial and feeders should be kept as low as possible, as each additional insulator provides another possible leakage path in parallel.

(*e*) LOSS DUE TO EDDY CURRENTS IN NEIGHBOURING CONDUCTORS.—Any metallic bodies near the aerial or feeder are in the direct inductive field, and will have currents induced in them. The magnetic field due to these currents reacts on the aerial by ordinary "transformer" action, and increases the loss resistance. For this reason the aerial and feeders should be kept as far away from metal as possible.

Similarly, the feeder should be at the maximum possible distance from, and never parallel to, stays, guys, etc. In spite of these precautions, currents will be induced in all the stays supporting the masts of a ship or shore station. These stays may be in one of three conditions :—

(i) Well insulated from earth, and split up into a number of well-insulated sections

(ii) Connected to earth by a low resistance path

(iii) Connected to earth by a high resistance path.

Metal stays or halyards should not be allowed to rub across each other, since this is a fruitful source of crackles in reception.

For both transmission and reception, condition (i) is best, then condition (ii) ; condition (iii) should never be tolerated.

If, therefore, the rigging insulators become so coated with dirt, paint, or tar, that they cease to insulate satisfactorily and cannot be cleaned, they should be short-circuited. Stays and shrouds should be carefully earthed at the point where they are connected to the hull.

When transmission is taking place in a ship, all the stays, etc., will be set in oscillation at a frequency depending on their length. Stays should, therefore, be divided up into sections of such a length that minimum interference is caused to reception of short waves.

At a shore station the best condition is that the masts themselves should be insulated from earth. This is not easy to achieve, however, for mechanical reasons. If the masts are of metal, they should be so placed that the feeder does not have to run near any of them.

This loss increases with the square root of the frequency, for the same reason as conductor losses.

25. **Effective Total Resistance.**—With a given aerial and a given current, we have seen that :—

(*a*) Radiation resistance varies directly as f^2.

(*b*) Resistance corresponding to conductor losses in the aerial and earth systems, and to eddy current losses, increases as f increases, and may be taken as proportional to $\sqrt{f}$.

(*c*) Resistance corresponding to dielectric absorption losses, leakage and brushing, decreases as f increases, and may be considered as inversely proportional to f.

These facts are conveniently summarised in Fig. 23. Three curves—(*a*), (*b*) and (*c*)—are drawn showing how the above losses vary with f. The dotted curve illustrating the loss resistance is obtained by adding the curves (*b*) and (*c*), and shows that at some point, such as X, the total loss resistance is a minimum. The top curve, showing the total effective resistance, is obtained by adding curves (*a*), (*b*) and (*c*).

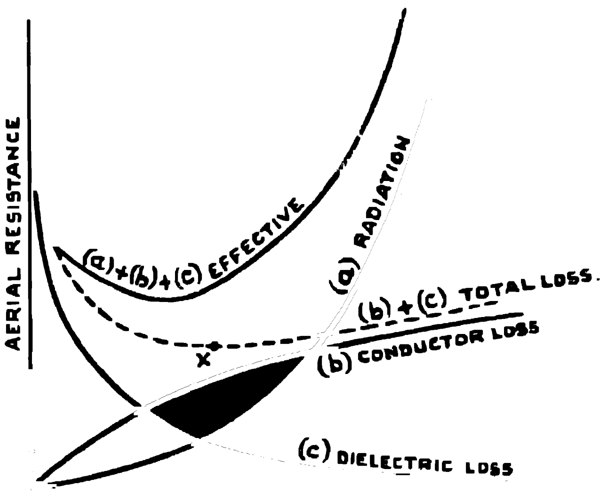
The curves given above are derived from theoretical considerations. It will be found, however, that the "effective resistance" curve of any aerial will have the same general shape as the one given.

26. **Aerial Efficiency.**—The reduction of loss resistance of aerial systems is especially important in high-power long-range shore stations, using low frequencies.

In the case of low frequencies it was seen in paragraph 11 that the radiation resistance is necessarily low. Aerial efficiency measures the ratio of power radiated to power supplied, and can be defined, therefore, as the ratio of the radiation resistance to the total aerial resistance.

$$\eta = \frac{\text{Power radiated}}{\text{Power supplied}} = \frac{R_r}{R_r + R_L}$$

where R_L denotes the loss resistance.



VARIATION OF AERIAL RESISTANCE WITH FREQUENCY.

FIG. 23.

Let l denote the aerial height. If the ratio of l/λ is kept constant as the frequency is increased, it will usually be found that η increases from very low values, often to values as high as 90 % for H/F. aerials; this is due to the decrease in R_L of the above equation as the physical length of the aerial becomes smaller.

For a given frequency, the only ways of increasing aerial efficiency are :—

(*a*) To increase the height, by erecting very costly masts which are in fact unpractical above about 300 metres, or

(*b*) To reduce the losses in the aerial and earth circuits, so that they are small compared with the radiated energy.

In the case of low frequency shore stations the special methods employed to keep the loss resistance down to very low values will depend upon the nature of the earth in use.

The case of a CONDUCTIVE EARTH is well seen at one station in France. The earth system there consists of a network of copper wires spaced 10 metres apart, divided into sections, and balanced electrically so that each section carries approximately the same current. The sections of the earth network should be symmetrical around the station, as regards both length and disposition.

The impedance of each of the earth sections is equalised with the help of artificial inductances and condensers.

The saving in energy loss may be appreciated by considering a numerical example. Let the resistance of each of the earth connections be R. Considering the simple case of only two earth leads, it will be assumed that, due to bad balancing, the current in one lead is twice that in the other. This means that two-thirds of the current will flow in one lead and one-third in the other and the total losses of the system may be written

$$(\tfrac{2}{3}I)^2 R + (\tfrac{1}{3}I)^2 R = RI^2 (\tfrac{4}{9} + \tfrac{1}{9}) = \tfrac{5}{9} RI^2.$$

Supposing, however, that the impedances of the two paths are equalised, the loss would be given by

$$(\tfrac{1}{2}I)^2 R + (\tfrac{1}{2}I)^2 R = \tfrac{1}{2}RI^2.$$

The energy loss is clearly smaller in this case.

The case of a COUNTERPOISE earth is illustrated in Fig. 24, which shows wires running radially outwards on insulated supports, without connection being made to earth plates at the outer end of the system. The function of the screen is to intercept the lines of force from aerial to earth, and to carry the return current on the screen wires rather than by the earth. Since no current flows in the earth itself, the losses due to poor conductivity are avoided. To get sufficient aerial capacity the surface area of the screen should be large. It should extend, on all sides, beyond the area covered by a plan view of the aerial system, by a distance equal to the height of the aerial.

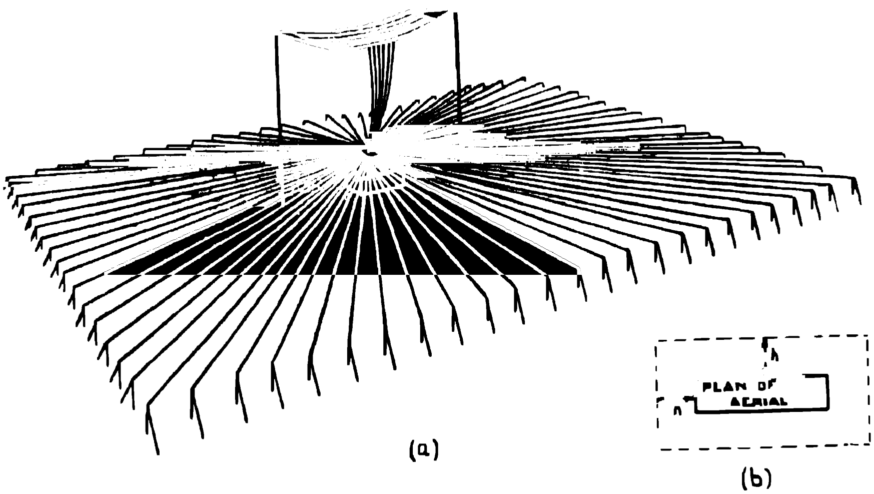


FIG. 24.

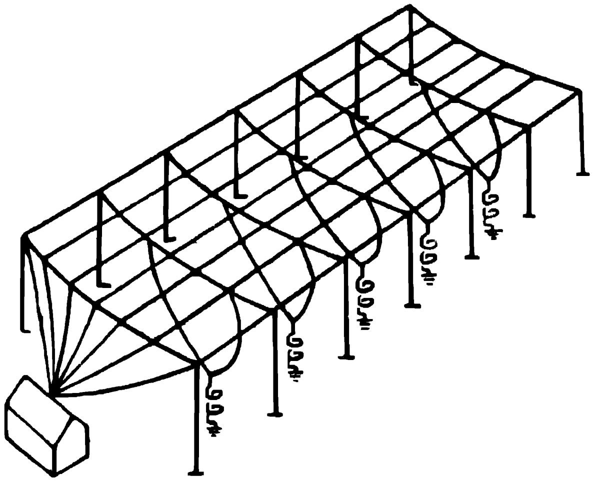
Loss resistance may also be reduced very much by the use of a MULTIPLE TUNED AERIAL —Fig. 25.

The system is due to Alexanderson, and is of particular value when the position of the station renders it difficult to obtain a good earth, *e.g.*, when it is built on stony or sandy soil.

It constitutes, in effect, a number of separate aerials in parallel, each with its own open-air tuning coil. Assume that the system consists of four aerials, each being tuned and having a current of I amps. If the earth resistance is R ohms, the total power lost will be $4I^2R$. If, however, only one aerial is employed, having the same earth resistance but with a current of 4I amps. in the aerial, the power lost would be $(4I)^2 R$; this is $16 I^2 R$, or four times the previous amount.

These aerials are expensive to construct and the process of changing the wave frequency is a slow one.

The conductive earth in modern warships is usually very good compared with that of the average shore station. The screening cases of receiving sets are bonded to the racks on which they are mounted and an effective earth is obtained when the latter are bolted down to the metal deck. Formerly it was the custom to provide special earth wires which passed up the inside of the aerial trunk and thence to the ship's side.



MULTIPLE TUNED AERIAL.

FIG 25.

27. **Ship Aerials.**—The arrangement of aerials in a warship presents problems not usually encountered elsewhere. With the exception of the largest liners, ships (and small shore stations) require to use one line of communication only; they are required to be able to receive on one frequency only at any given time, and in general to be able to transmit on the same frequency. Large shore stations may have to be able to receive and transmit on many frequencies simultaneously. To make this possible, receivers and transmitters are spaced by a distance of several miles, the latter being controlled from the receiving station. It is essential to have this separation in order to remove the receivers from the swamping influence of the transmitters.

It is necessary for a warship to be able to receive on a number of different wave frequencies, and it is very desirable that reception should not be subject to interference due to transmitters in the ship working on other frequencies. It is obviously impossible to separate the transmitting and receiving aerial systems by any appreciable distance. There are, however, two methods of minimising this mutual interference; they are termed respectively the " Spaced aerial system " and " Central control system."

THE SPACED AERIAL SYSTEM.—This system consists in placing transmitting and receiving aerials as far apart as possible, *i.e.*, at opposite ends of the ship.

Fig. 26 shows a ship fitted with the spaced aerial system. The main transmitting aerial is

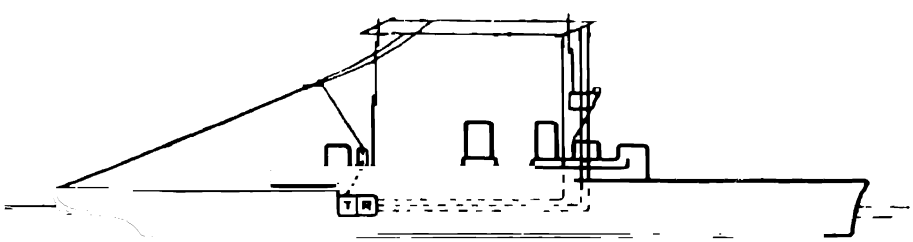


FIG. 26.

rigged between yards on the foremast and the mainmast, the end being led down to the transmitter in the wireless office. The receiving aerials are simple vertical aerials hung in the form of a curtain before the foremast, and are connected by transmission lines to the receivers in the wireless office. (For **transmission lines,** *see* paragraph 36.)

THE CENTRAL CONTROL SYSTEM.—The aerial rigging is much the same as in the case of the alternative system, but the receiving aerials are directly connected to the receivers located in a special compartment in the fore part of the ship.

Fig. 27 shows a ship fitted with the central control system. The transmitter is controlled from the receiving room by controller lines, as in the case of shore stations with spaced transmitters and receivers. In R.M.S. " Queen Mary " the receiving station is 400 ft. from the transmitters.

It should, however, be noted that the separation of transmitters and receivers is a disadvantage from practically all points of view, except as regards the abolition of interference between transmitters and receivers. It is undesirable to separate operators from the transmitters, but where this is done, the first requirement becomes a somewhat complicated control system, the object of which is to place each operator in as close touch as possible with his own transmitter.

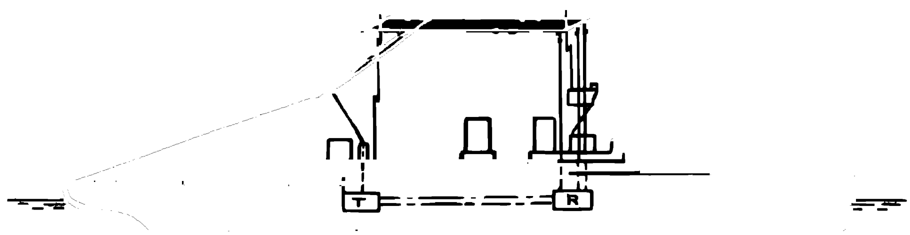


FIG. 27.

It has been observed that the aerial rigging is much the same in the two cases. The receiving aerials are to some extent screened from the transmitting aerials by the foremast, which, in a warship usually carries large metal masses constituting bridges, spotting tops, control towers, etc.; from the point of view of incoming signals this screening is undesirable. They are also spaced by as great a distance as possible from the vertical portion of the main transmitting aerial; this is especially advantageous, since the latter is the part of the aerial associated with the most intense field Interference is still further minimised by terminating the roof portion of the transmitting aerial at some distance from the foremast; in the figures, the insulators mark the end of the main transmitting aerial.

Where two or more transmitters are installed in a ship it is advisable to fit them in the same or adjacent compartments, so as to keep their aerials remote from the receiving aerials.

A further contribution to the solution of the problem of simultaneous transmission and reception in the same ship may be made by fitting **R/F filters** between the aerials and the receivers which they supply. This is a matter which might introduce considerable complication into a receiver outfit, and, in some of the larger ship installations in liners, the problem has been approached by fitting two separate exchanges, one for the various receiving aerials and one for the filter units. Filters may be connected to the aerial exchange so that any receiver can be connected through any filter to any aerial. Such an arrangement might also make it possible to supply several filters from one aerial and several receivers from one filter. The design of such R/F filters must depend upon the cause of the interference which it is desired to suppress, and it is here assumed to be due to a powerful transmitter working on a frequency widely separated from that of the signal which it is desired to receive. In actual naval practice, the frequencies on which a warship will be required to transmit or receive are fixed whilst the ship is serving on one station; moreover, on a change of station, the new frequencies required will not be far removed from those previously used. It therefore appears

possible to divide the total frequency range used in Naval communications into a series of sections, and say that, in general, one transmitter and one receiver will be working in each section. If a band pass filter covering one of these sections is fitted between an aerial and a receiver, it should be theoretically possible to ensure that no transmitter working outside this section will interfere with the receiver. Such a band pass filter would consist of a chain of coupled circuits, but it could be arranged so that the circuits were all pre-set, requiring no tuning by the operator whatever the tuning of the receiver provided it be within the pass band of the filter. The response curve of such a band pass filter is one that shows a very high attenuation for signals outside the band and a low attenuation for signals inside it.

In general, filters for this use may be of two distinct types—"**band pass filters**" and "**tuned filters.**"

Band pass filters can be arranged to pass to the appropriate receiver all frequencies within a certain band, and reject all frequencies outside it. Tuned filters can be arranged to pass a very narrow band of frequencies. Band pass filters have the advantage of easy operation, but suffer from the disadvantage that any interfering signal within the band is passed on to the receivers; this interference may be due to one of the harmonics of some transmitted frequency, in which case its presence would be serious. Tuned filters, on the other hand, could protect the receivers from all interference not at nor very close to the signal frequency, therefore offering a much greater degree of protection, though requiring careful re-tuning for any change of frequency.

There are many different varieties of filter sections which could be utilised, but for the purpose of rough classification the following types may be noted :—

(*a*) Filters with acceptor circuit characteristics connected in parallel between aerial and receiver.

(*b*) Filters with rejector circuit characteristics connected in series between aerial and receiver.

(*c*) Filters with rejector circuit characteristics, connected in parallel between aerial and receiver, with a high resistance in series with each parallel circuit.

A further considerable decrease in the amount of interference could be obtained by a reduction of the amount of power radiated as harmonics by the ship's transmitters. This could be effected by increasing the number of coupled circuits. That would involve either an increase in the space occupied by a given transmitter, or, alternatively, a reduction in the power used, and in either case some complication would be added to the transmitter, due to the additional tuned circuits. In R.M.S. "Queen Mary," where every attempt has been made to achieve simultaneous transmission and reception, the transmitting frequencies have been chosen so that none of their harmonics fall within 10 kc./s. of the frequencies in use for reception.

Transmitting aerial design is subject to grave restrictions imposed by the position of the funnels, stays, guys, etc. The principal object in fitting the main L/F aerial should be to make its radiation height as great as possible. A "**flat roof**" form of aerial is commonly used ; it consists of four or more parallel wires spaced by spreaders at either end. This is a simplification of the older Naval aerials that consisted of cylindrical arrangements of wires in parallel, the well-known "sausage" aerials. It is undesirable that the free end of an "L" aerial should be at a lower level than the point at which the roof joins the feeder down lead, for in that case the field due to the current in the roof has a component in opposition to that produced by the down lead.

In a ship it is difficult to make provision for special H/F transmitting aerials, and usually the main aerial is used in one of two conditions. It may be operated as an untuned aerial, or as a tuned one. Using it as an untuned aerial, it is equivalent to a resistance for the reason explained more fully in paragraph 15. Using it as a tuned aerial, adjustments are made until the electrical length of the aerial is a multiple of the wavelength in use. When this is the case, standing waves will be set up, and the aerial system will be operated as the electrical equivalent of a $\lambda/4$ aerial.

28. Field Strength (X) at a Point within Range of the Direct Ray.—In paragraph 7 the

general expression for the instantaneous electric field " x " at any distance " d " for which the inductive fields may be neglected, is quoted as

$$x = \frac{2\pi f \,.\, \delta l}{d} \mathcal{J} \cos \omega \left(t - \frac{d}{c}\right) \cos \theta.$$

Considering only the amplitude part of this expression, and omitting the factor " cos θ ", which expresses the distribution of energy in the vertical plane—

$$\mathfrak{X} = \frac{2\pi f \delta l \mathcal{J}}{d}$$

or in R.M.S. values, using I_s to denote the current in the " sending " aerial, and substituting for in terms of λ

$$X = \frac{2\pi f \delta l I_s}{d} = \frac{2\pi c \delta l I_s}{d\lambda}.$$

This assumes that the current I_s is uniform over the length δl of the radiator. Now it was seen in paragraph 10 that for any aerial we can derive an effective height, over which the current is constant. If this is l_r for an earthed $\lambda/4$ aerial, the effect of the " image " is to produce a radiator of apparent length $2l_r$.

The factor 2 is dependent on the assumption that the earth is a perfect conductor: where this assumption is a bad one (*cf.* paragraphs 16 and 17), the factor 2 is too big.

Writing $2l_r$ for δl we have

$$X = \frac{2\pi c 2 l_r I_s}{d\lambda} = \frac{4\pi c l_r I_s}{d\lambda} \quad \ldots\ldots \quad \text{in E.M.U's.}$$

Substituting for c and expressing X in practical units $\left(\frac{\text{volts}}{\text{cm.}}\right)$, and measuring I_s in amps., we have

$$X = \frac{4\pi \times 3 \times 10^{10}}{10^8 \times 10} \frac{l_r I_s}{\lambda d} = \frac{377\, l_r I_s}{\lambda d} \quad \ldots\ldots \quad \text{in } \frac{\text{volts}}{\text{cm.}}$$

if λ, l_r and d are in cms., or

$$X = \frac{377\, l_r I_s}{\lambda d} \quad \ldots\ldots \quad \text{in } \frac{\text{volts}}{\text{metre}}$$

if λ, l_r and d are in metres, and similarly for other units of length.

This is a well known and widely used formula, which gives the approximate field strength to be expected at distances exceeding 5λ and up to 200 to 300 miles, when $l_r I_s$—the **metre amperage** of the transmitter—is known. It makes no allowance for attenuating losses (absorption losses) suffered during propagation, and for long distances Austin and Cohen have added an exponential factor $e^{\frac{ad}{\sqrt{\lambda}}}$

where " e " is the base of a Naperian logarithm,

where " a " is an experimentally determined constant which varies with the time of the day and the path followed, and is approximately 0·0015 for transmission over sea ; and

where " d " and λ are both measured in kilometres.

The formula then is

$$X = \frac{0{\cdot}377 \times 10^6\, l_r I_s e^{-\frac{ad}{\sqrt{\lambda}}}}{\lambda d} \quad \ldots\ldots \quad \frac{\text{micro volts,}}{\text{metre}}$$

where d, λ, and l_r are measured in kilometres.

The attenuation factor shows clearly that as f increases, the value of X at a given distance rapidly falls, and the "**ground ray**" virtually disappears at a short distance from the transmitter.

The fact of long range H/F propagation in spite of the disappearance of the ground ray, led to the discovery of the reflecting properties of the ionosphere, and the extensive use of the "**sky ray.**" The Austin Cohen formula cannot be applied to the indirect ray, which is relatively less attenuated than the direct ray. This subject is discussed more fully under the heading "Propagation of electromagnetic waves." Neither the above classical Austin Cohen formula, nor the many more modern variations of it, are capable of giving a complete account of the facts. Only one other example will be given; as a result of measurements made in 1922 between Horsea Island and H.M.S. "Antrim" on a cruise to Sierra Leone, the following formula was suggested by H.M. Signal School:

$$X = \frac{0 \cdot 377 \times 10^6 \, l_r \, I_s \, e^{-\frac{0.0014d}{\lambda^{1.6}}}}{\lambda d} \quad \ldots\ldots \quad \frac{\text{micro volts}}{\text{metre}}.$$

29. **Voltage Produced in a Receiving Aerial.**—Transmitting and receiving aerials have properties which are reciprocal in nature. Generally speaking, good radiating aerials also make good receiving ones. A receiving aerial serves, as it were, to make use of the alternating P.D. which exists between two points vertically spaced in the electrostatic field of the passing vertically polarised radiation.

If the effective height of the receiving aerial is l'_r, the R.M.S. voltage in the aerial will be given by

$$V = Xl'_r = \frac{377 \, l_r \, l'_r \, I_s}{\lambda d}.$$

Receiving aerials are treated in greater detail at a later stage.

30. **Typical Values of the Field Strength (X) in Micro-volts per Metre.**

Frequency.	Field strength µV per metre.	Remarks.
M/F Broadcast . . .	100,000	Very close to high power stations
.. ..	5,000–30,000	Requisite value for good telephony service in towns where "man-made static" is bad.
,, ...	10,000–15,000	Limits of field strength produced by Droitwich B.B.C. station at Portsmouth
,, ...	1,000	Quite a good telephony signal, but needs H/F amplification, poor "signal to noise" ratio.
,, ...	500–5,000	Average signals at 1,000 miles at night from a high power broadcast station.
,, ...	100	A readable telegraphy signal; needs a good receiver; much interference.
,, ...	50	Often considered the minimum satisfactory signal.
,, ...	10	Extremely weak; needs a receiver with many refinements.
H/F Broadcast and Beams ..	100	Quite a strong telephony signal; uses the indirect ray.
,, ,, ..	10–20	Satisfactory reception.
,, ,, ..	1·0	The minimum acceptable signal to the G.P.O. in their radio telephony services.

It will be noted that a much smaller signal gives satisfactory service at high frequencies. The reason for this is that atmospheric noises at H/F are much less intense than at L/F and M/F. It is accordingly possible to preserve an adequate "signal to noise" ratio, using a much smaller signal strength. For this reason also, it is clearly impracticable to give comparative values of X in terms of any R code.

31. Transmitting Aerials—Numerical Exercises.

(a) A transmitting aerial having an effective height of $2/\pi$ times the physical height has a current at the base of 725 amps., at a frequency of 16 kc./s. ($\lambda = 18,800$ metres) Given that the physical height is 175 metres, and the aerial efficiency is 11 per cent find—

(i) the value of X and of H at 175 kilometres ;

(ii) the value of R_r ;

(iii) the power radiated ;

(iv) the power in the aerial ;

(v) the R.M.S. value of the voltage produced in a receiving aerial of 100 metres effective height at 175 kilometres

(1) Effective height $= \frac{2}{\pi} \times 175 \doteq 111 \cdot 3$ metres.

$$X = \frac{377 \times 111 \cdot 3 \times 725}{18800 \times 175000} \frac{\text{volts}}{\text{metre}} = 9 \cdot 25 \frac{\text{millivolts}}{\text{metre}}$$

$$= 9 \cdot 25 \times 10^{-5} \frac{\text{volts}}{\text{cm.}}$$

(para. 2) $$H = \frac{X}{300} = \frac{9 \cdot 25 \times 10^{-5}}{300}$$

$$= 3 \cdot 08 \times 10^{-7} \frac{\text{lines}}{\text{cm}^2} \quad \ldots\ldots\ldots \quad \left(\text{or } \frac{\text{dynes}}{\text{pole}}\right).$$

(2) $$R_r = \frac{1580 \times 111 \cdot 3^2}{18800^2} = 0 \cdot 0554 \text{ ohms.}$$

(3) Power radiated $= 725^2 \times 0 \cdot 055 = 29$ kW.

(4) Power in the aerial $\frac{29 \times 100}{11} = 264$ kW.

(5) $V = 9 \cdot 25 \frac{\text{mV}}{\text{metre}} \times 100 \text{ metres} = 0 \cdot 92$ volt.

(*b*) On 100 kc./s. ($\lambda = 3{,}000$ metres) the field at 1,469 kilometres from Horsea Island W/T station was observed to be 86 μV per metre. Assuming the " metre amperes " of the Horsea aerial to be " 100 × ·35 ", using the Austin Cohen formula, obtain a " calculated " value for X at the above distance.

$$X = \frac{0{\cdot}377 \times 10^6 \times 0{\cdot}1 \times 35}{3{\cdot}0 \times 1469} e^{-\frac{0{\cdot}0015 \times 1469}{\sqrt{3{\cdot}0}}}$$

$$= 298\, e^{-\frac{2{\cdot}2}{1{\cdot}73}} = 298\, e^{-1{\cdot}27}$$

$$= 298 \times 0{\cdot}28 = 83{\cdot}5 \ \frac{\mu V}{\text{metre}}.$$

This result is in accordance with the commonly observed fact that the actual field strength at a given distance from a transmitter is usually bigger than that calculated from the Austin Cohen formula.

32. **Aerial Measurements.**—In the foregoing discussion the radiation resistance of an aerial was described as fictitious, and the effective aerial height has been shown to be not identical with the physical height and difficult to compute. It is therefore very reasonable to wonder how aerial constants can be obtained with some precision; several of the fundamental properties of aerials are made clearer by considering this matter briefly.

Effective Total Aerial Resistance. This has been shown to be a composite quantity; it can be measured in various ways :—

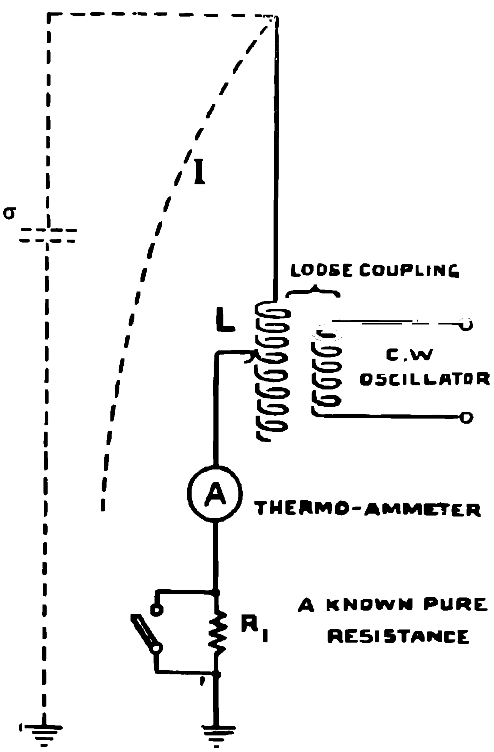


FIG. 28.

(*a*) THE RESISTANCE VARIATION METHOD. Fig. 28 represents an equivalent $\lambda/4$ aerial, having the aerial ammeter A in the earth lead, together with a known resistance R_1 which must be non-inductive and non-capacitive. A switch is shown by which R_1 may be introduced into the circuit or cut out at will. L represents a small coil, to which is coupled a C.W. oscillator. R_L will be used to denote the effective total aerial resistance. Using a loose coupling between the oscillator and the aerial, the latter should be carefully tuned to resonance, with resistance R_1 short circuited. Let the aerial current in this case be I_1. The pure resistance R_1 is switched into the circuit, and the aerial current falls to some other value I_2. It follows that, since the aerial is resonant, we may write

$$I_1 = \frac{E}{R_L}, \quad \text{and} \quad I_2 = \frac{E}{R_1 + R_L}$$

$$\text{Hence} \quad \frac{I_1}{I_2} = \frac{R_1 + R_L}{R_L}.$$

Since R_1 is a known resistance, it is clear that R_L can be calculated. For various reasons it is not a very accurate method.

(*b*) THE SUBSTITUTION METHOD.—Fig. 29 represents an equivalent $\lambda/4$ aerial which —with the change-over switch in one direction—has its circuit completed to earth through the ammeter A and coupling coil. With the change-over switch in the other direction, an **artificial aerial** or substitution circuit takes the place of the actual aerial. With a given input from the C.W. oscillator, the constants of the artificial aerial circuit should be adjusted to give the same aerial current as that given by the actual aerial. The resistance of the substitution circuit must be non-inductive and non-capacitive, and the capacity itself should be as pure as possible. If these instruments are calibrated, it follows that both aerial resistance and aerial capacity are obtained by direct reading. This method of measuring total aerial resistance is a very satisfactory one, but requires slightly more elaborate apparatus than the resistance variation method.

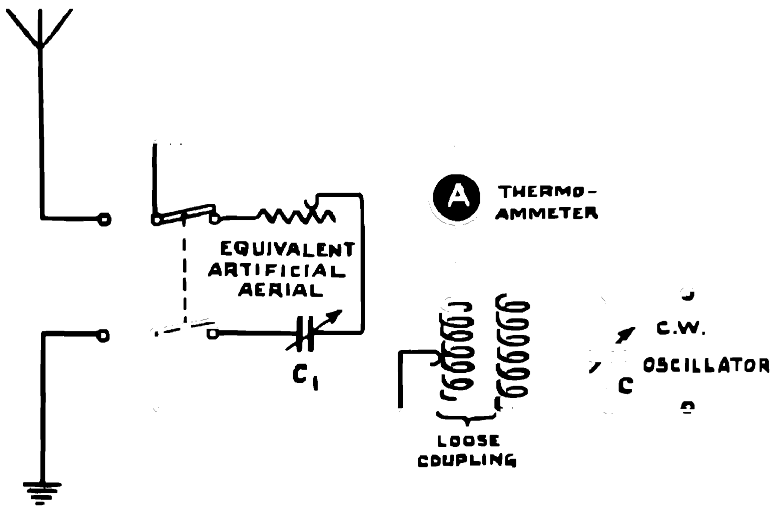


FIG. 29.

(*c*) THE AMMETER AND VOLTMETER METHOD.—Fig. 30 (*a*) and (*b*) represents a $\lambda/4$ and a $\lambda/2$ aerial which could be excited respectively in the manner shown.

IN THE CASE OF THE $\lambda/4$ AERIAL, the circuit at the base is tuned to the frequency of the C.W. oscillator, the aerial being said to be in tune when the ratio V/I is a minimum. The value of this ratio represents the total aerial impedance, which at resonance is a pure resistance.

IN THE CASE OF THE $\lambda/2$ AERIAL, the circuit at the base is similarly tuned until the ratio V/I_a is a maximum. The value of this ratio gives the total aerial resistance at the base of a $\lambda/2$ aerial. In this case, since a current anti-node occurs at the centre, it follows that when the ammeter I_a indicates a minimum the ammeter I will indicate a maximum.

Instead of considering the base resistance of a $\lambda/2$ aerial, it is more useful to refer the resistance to the aerial current anti-node at the centre, for this gives a figure for the total aerial resistance which is comparable in meaning with that of the $\lambda/4$ aerial. It is given by—

$$I^2 R = V I_a$$

or

$$R = \frac{VI_a}{I^2}$$

where $R = R_r + R_L$ .... (compare paragraph 26).

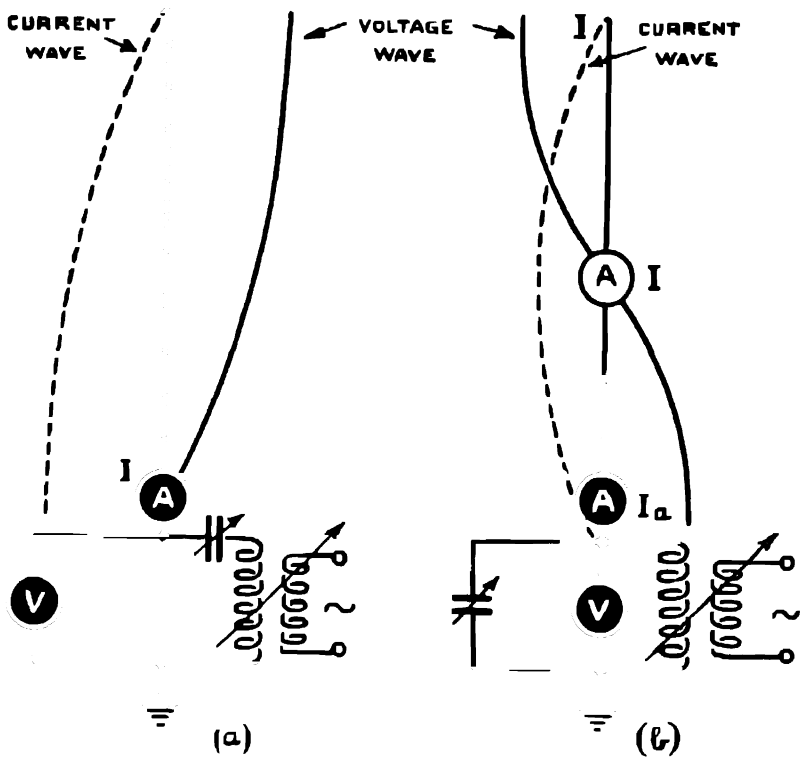


FIG. 30.

These measurements may be made clearer by considering possible numerical values of the two total resistances ; it is possible that radiation resistance might form 80 per cent in the case of the λ/4 aerial, and possibly slightly more in the λ/2 case.

λ (metres).	Type of aerial	V (volts)	I (amps).	I_a (amps)	Power in aerial (W) Watts	Total resistance (ohms).
285	$\frac{\lambda}{4}$	100	2·5		VI 250	$\frac{V}{I} = 40$
285	$\frac{\lambda}{2}$	1300	2·0	0·4	VI_a 520	$\frac{VI_a}{I^2} = 130$

With the λ/2 aerial, it is, naturally, somewhat inconvenient to have to read an ammeter situated half-way up an aerial. In one case during some B B C. experiments, conducted in 1929, on an aerial which was held aloft by means of a captive kite balloon, it was arranged that the ammeter I in the diagram should be a large one, so that it could be read with the help of a telescope located on the ground. It will be noted that the base resistance of the λ/2 aerial is about 3,250 ohms.

This figure only gives an approximate indication of the " order " of this base resistance ; in practice it is influenced by a number of factors, such as the proximity of the aerial to earth, and the nature of the aerial itself. Tubular or multi-wire aerials would not give the same value as a single wire λ/2 aerial.

EFFECTIVE AERIAL HEIGHT (l_r).—It has been seen that the effective height of an aerial is intimately related to the distribution of current in it. By making assumptions about this distribution it is possible to arrive at an estimate of aerial effective height in simple ideal cases. The practical determination of effective height is achieved by measuring the field strength (X) at a point distant about 5λ from the transmitter. We have the Hertzian formula

$$X = \frac{377\, l_r\, I_s}{\lambda d} \quad \ldots\ldots \quad \text{Hence } l_r = \frac{X \lambda d}{377\, I_s}$$

The methods used in the accurate determination of X, upon which the measurement of l_r depends, will vary to some extent with the magnitude of the field strength to be measured.

The signals are usually received on a frame aerial tuned by a condenser and having a thermo-milliammeter in circuit, Fig. 31.

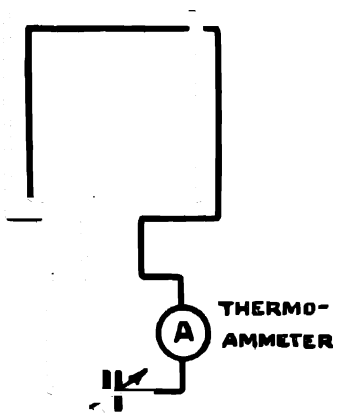


FIG. 31.

The principle of the frame aerial is treated fully in the section on Direction Finding, where it is shown that an incoming signal produces a loop E.M.F. given by the expression

$$E_L = \frac{2\pi X A}{\lambda} \quad \ldots\ldots \quad \text{(where A = area of the loop).}$$

This loop E.M.F. will produce a resonant current determined by the resistance of the circuit and given by $E_L = IR$. We thus have

$$E_L = IR = \frac{2\pi X A}{\lambda}.$$

Where the field strength is large, it is convenient to measure I; where the field strength is small, it is usually more convenient to measure E_L by means of a previously calibrated valve voltmeter. In either case X is calculated from the above formula.

With the knowledge of the effective height, a value may readily be obtained for the radiation resistance (R_r), and the power radiated.

POWER IN THE AERIAL.—In many cases the power of a transmitter is described in terms of the total power in the aerial (paragraph 10). This can be measured by the same apparatus that was described in the "ammeter and voltmeter method" of determining total aerial resistance.

In the case of the λ/4 aerial, Fig. 10 (*a*), the power at resonance is given by VI; it is also, of course, given by I^2R, where R denotes the total aerial resistance. In the case of the long wave B.B.C. aerial at Droitwich, the aerial current is approximately 86 amperes, and the total aerial resistance is approximately 20·4 ohms. This gives a power in the aerial of approximately 150 kW.

In the case of the λ/2 aerial, the power at resonance is given by VI_a; it is also given by I^2R, where R = total aerial resistance.

With a knowledge of the power in the aerial and the power radiated, we can arrive at an estimate of the aerial efficiency.

33. **Scale Models.—The Principle of Similar Aerials.**—In the design of big aerial systems, it is not found possible satisfactorily to predict all of the constants. It is necessary to conduct small scale experiments, usually on the actual site.

Small scale experiments in mechanical engineering are commonplace; stresses in bridges are investigated by using models. The counterpart of the wave distribution produced by a big liner moving at a certain speed through water can be seen by towing a scale model at a different speed in

a tank. The underlying mechanical principle is the "**Theory of similar structures,**" the results of which apply equally to aerial design. For any simple aerial, the field strength at a distant point is given by

$$X = \frac{377\, l_r I_s}{\lambda d}.$$

At a given distance d, the same value of X can be produced by two different aerials on different frequencies, provided that $\frac{l_r I_s}{\lambda}$ is maintained constant.

Any two transmitters having the same value of $\frac{l_r I_s}{\lambda}$ may be defined as "**similar transmitters.**"

The present B.B.C. medium frequency (long wave) aerial at Droitwich is a single wire "T" aerial, separated by two 700-ft. insulated masts. Before the erection of these masts, experiments were conducted using models to 1/10th and 1/7th scale. The 1/10th scale model used 70-ft. masts, a frequency corresponding to a wavelength of 150 metres, and a given current. Experiments at that frequency proving satisfactory, it was considered sound to erect 700-ft. masts to work at a wavelength of 1,500 metres.

It is usually safe to say that a scale model of an aerial will have the same radiation resistance and directional properties as the full-sized aerial. This principle, that an aerial worked at one frequency is equivalent to a larger one worked at a lower frequency, is of great assistance to the radio engineer.

34 **Masts for Shore Stations.**—For transmitting purposes the design of masts depends very much on their height, which in turn depends upon the frequency on which it is desired to transmit; H/F transmitters usually do not require masts as high as those necessary for the transmission of low and medium frequencies. The design of a mast also depends on the function it has to perform. In some aerial systems the masts simply serve to support wires which are slung between them; these wires are called *triatics*. A triatic may form a part of the aerial or radiating wire, or it may serve simply as a means of suspension for the aerial system. In other cases, the mast may support a vertical aerial, or it may itself be operated as a vertical radiator. Masts are expensive to erect and to maintain, and the introduction of $\lambda/2$ vertical radiators for M/F broadcast work up to a physical height of 800 ft. using one mast only, produces a considerable economy. In America, a single guyed vertical radiator 500 ft. in height would cost about £4,000. The B.B.C. station in North Ireland provides one example of this practice. In the case of M/F broadcasting stations, it should be stated that there are also other advantages attached to the use of a single mast and vertical radiator. Prominent among these are—

(*a*) The horizontal polar diagram is rendered more symmetrical by the elimination of interference shadows produced by masts. This kind of trouble was particularly acute at the older Daventry broadcasting station, where a conventional type of aerial was used and the masts were non-insulated. The production of shadows due to the absorption of energy by masts is much reduced by insulating them, a practice which is now usually considered imperative. The B.B.C. difficulty was partly overcome by detuning the masts by hanging "tail-wires" from the cross arms.

(*b*) The vertical $\lambda/2$ type of radiator has been found to increase the area over which *reception free from fading* can be achieved. It does this by virtue of the fact that it increases the amount of low angle radiation while reducing the amount of high angle radiation; the interference effect usually known as "**fading**" will therefore be minimised, due to a reduction in the amount of down coming radiation.

Masts are often seen to be surmounted by cantilever arms, usually for the purpose of suspending other triatics which may in turn support curtains of **reflecting aerials.**

For the purposes of classification, masts may be divided into two types, the guyed type and the self-supporting type.

THE GUYED TYPE.—This may be sub-divided into—

(a) Tubular steel masts, and

(b) Wooden or steel lattice masts.

The tubular steel type is suitable for short masts and is simple to erect. On one occasion, one of H.M. Ships had to improvise a mast at short notice; it was found possible to set up a 60-ft. mast of that style, using lengths of galvanised water tube as the raw material. The water tubes were in 15-ft. lengths, 1½-in. internal diameter, one end of each being provided with a screw thread and the other a tapped ferrule. Four of these lengths were screwed together to make one 60-ft. length. At each joint, a piece of 1½ ins. (external diameter) tube, 3 ft. in length, was driven in to act as a stiffener, bearing in mind that the joints are the weakest portions of such a mast. A wooden block, with a 2-in. hole drilled in it, was provided as a heel fitting, and the top of the mast was provided with a wooden cap to prevent accumulation of rain water within the mast. Various blocks and endless halyards were also provided at the top, and the mast was supplied with three sets of four stays at heights of 20 ft., 40 ft., and 60 ft. from the heel. The stays were of 1¼-in. flexible steel wire rope. The mast was raised into position by means of a sounding boom rigged as a derrick, the stays being carefully tended during the operation Twelve hands erected the mast in 20 minutes. Subsequently, care was taken to remove traces of bending in the mast by adjustment of the stays. The mast withstood an aerial pull of 150 lbs. normal to the masthead, and was unaffected by the most severe gales. It was thought that, in an emergency, there would have been no constructional difficulty in erecting a 100-ft. mast of this type by simply providing extra stays.

The new B.B.C. station at Droitwich employs two stayed lattice steel masts, 700 ft. high and 600 ft. apart, the section of the masts being triangular and each being provided with an electrically operated lift within the structure. Each mast is supported by three sets of three stays, and the masts are insulated at the base by porcelain insulators capable of withstanding an electrical stress of 7,000 volts at 200 kc./s., and a mechanical stress of about 150 tons. These masts serve to support a single wire " T " aerial for the transmitter working on 200 kc./s., and also a $\lambda/2$ aerial, 450 ft. long, which is hung from a triatic slung from the top of one of the masts.

THE SELF-SUPPORTING TYPE.—Within limits, lattice aerial masts may be made self-supporting by splaying out the base upon which they stand. This type of mast is useful in cases where it is desired to save the ground space which a stayed mast would need. It is a useful way of supporting aerials on the tops of high buildings. These aerial masts are usually more expensive to construct than the guyed type. A service example is provided by the aerial "**arrays**" at Horsea Island. The array curtains are suspended from triatics slung between the cantilevers of five self-supporting steel towers, of height about 180 ft.

The transmission of low and medium frequencies necessitates the use of high and expensive masts, to support aerials of the loaded $\lambda/4$ type.

For the broadcast transmission of medium frequencies, evolutionary changes since 1929 have extended the non-fading "**service area**" and have popularised the **half wavelength type of aerial** using the single vertical radiator referred to above. The best British example is the North Ireland B.B.C. transmitter opened in March, 1936. In this case, the mast is a cigar-shaped lattice steel structure of height 475 ft., the steel structure itself being used as an aerial. The mast is stayed by three sets of insulated stays, and is supported on a ball and socket joint insulated from earth. The whole is surmounted by a 75-ft. topmast of adjustable height, carrying at its summit a horizontal ring 20 ft. in diameter. The sliding topmast was provided in order to have some means of adjusting the length of the aerial in case any change of the allotted frequency of the transmitter should take place. A 20-ft. ring, in appearance like a cartwheel, provides a concentrated capacity, sometimes called the "**capacity hat,**" and is associated with a small concentrated inductance. By proper

adjustments of the localised capacity and inductance at the top of the aerial, the form of the standing current wave may be varied at will, and it is found possible to—

(*a*) increase the radiation height of an aerial of given length, and

(*b*) control the form of the vertical polar diagram, adjusting it to give the maximum amount of low angle radiation, by making it electrically equivalent to a $0 \cdot 58\lambda$ aerial [*cf.* paragraph 40] when its physical height is less than $0 \cdot 5\lambda$.

In practice, the effect of the capacity hat is to reduce the length of the topmast necessary by about 25 ft. in the case of the North Ireland B.B.C. station. In general, the capacity hat may have other forms and may appear as a sphere, a cylinder, or a disc. In the two former cases, the concentrated inductance could conveniently be accommodated within the sphere or cylinder.

For the purpose of *reception*, aerials for all frequencies may be much shorter in comparison. It is only for economical or high-speed point-to-point working, or the reception of very weak signals, that elaborate aerial arrays become necessary.

35. **Coupling the Aerial to the Transmitter, Impedance Matching.**—It is a well-known principle in electrical engineering that to obtain maximum power dissipation in some impedance connected to a power unit, the value of the impedance should be equal to the internal impedance of the power unit.

In the simple case of a resistance R joined across the terminals of a battery of E.M.F. given by E, and with internal resistance b, the power dissipated in the resistance is given by

$$P = I^2 R = \left(\frac{E}{R + b}\right)^2 R.$$

It is easy to show mathematically that P will have a maximum value when $R = b$.

The applications of this principle are very numerous, and the following are some of the better known examples :—

(*a*) Matching the loudspeaker impedance at 800 cycles to the A.C resistance of the triode output valve or valves of a broadcast receiver. For maximum power, the output impedance of any triode valve in an amplifier should be similarly designed.

(*b*) Matching the oscillatory circuit impedance in a valve oscillator or transmitter to that of the valve. This is usually achieved by adjustment of the ***anode tapping point*** (A.T.P.) on the inductance constituting part of the oscillatory circuit. It should be noted that the oscillatory circuit may, possibly, not be coupled to an aerial, in which case the radiation and other losses will be very small. The LC circuit is then like a flywheel of an engine on no load, which is a reservoir of energy often called a " **tank circuit.**"

(*c*) Matching the transmitting aerial to the tank circuit, in order that the latter should present a resistance to the aerial equal to the total aerial resistance. This is achieved by adjustment of the aerial tapping point on the tank circuit tuning inductance, when the aerial is located in the immediate neighbourhood of a transmitter and is directly coupled to it. It therefore follows that in this case the tuning inductance will carry both the aerial and anode tappings, and thus acts as an auto-transformer.

(*d*) Matching the total aerial resistance, in the case of receiving aerials, to the input impedance of a receiver. This is usually done by means of a transformer coupling.

When the aerial is not located in the immediate neighbourhood of the transmitter or receiver, it will be necessary to join the aerial system to the transmitter by means of a connecting line, known as a " **transmission line.**" In the case of shore stations it is usually inconvenient, if not actually impossible, to locate aerials or aerial arrays in the immediate neighbourhood of the transmitter or

receiver, and it is usual to remove the aerial system to a suitable site which may be many wavelengths distant from the apparatus.

In joining up an aerial system to a transmitter by means of a transmission line, the matching referred to above has to be maintained, but a complication arises due to the special properties of the connecting line.

36. **Transmission Lines—Characteristic Impedance.**—A connecting line, made up of two parallel wires, has a certain capacity C per unit length and a certain inductance L per unit length, and the line is said to have distributed capacity and inductance. Fig. 32 (*a*) represents a feeder line of finite length; one end is joined to a source of supply and the other end is open. In many respects it is like an aerial, with the noticeable difference that the constants are more uniformly distributed. Travelling waves proceed towards the open circuited end where they are reflected, and by interference with the outgoing travelling waves the usual *stationary waves* are set up. In paragraph 18 it was observed that the phenomenon of stationary waves could also be explained in terms of "line resonance." A voltage anti-node, in a standing wave system, is to be regarded as precisely similar to the large resonant voltage obtainable across a condenser in a series A.C. circuit.

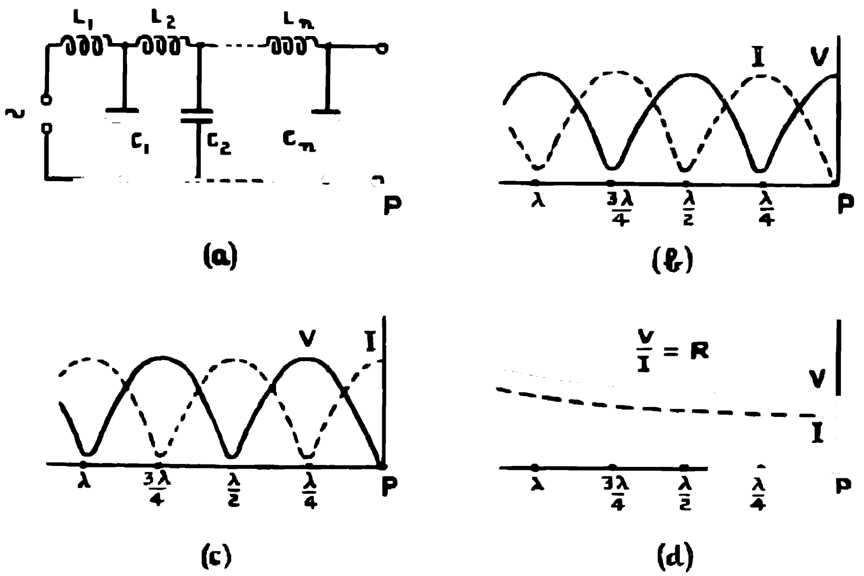


FIG. 32.

In the case of an aerial, standing wave diagrams were drawn in a conventional manner in Figs. 12, 13 and 14. Fig. 32 (*b*) represents the standing voltage and current waves on a feeder system open circuited at the *receiving end* "P." In this figure, standing waves are represented by an alternative method, which has the advantage of showing clearly that the nodes of current or voltage are imperfect ones. This advantage is offset by the failure to show the change of phase which takes place at points distant λ/2 apart; in this respect Fig. 14 is better. With reference to the receiving end at P, it may be assumed that the source of supply is located anywhere to the left of it. For example, at a distance from P equal to λ/4, the form of the voltage and current standing waves is that which already has been described in paragraph 16.

Standing waves must also be set up if we short-circuit the ends of the feeder lines remote from the source of supply. It may be considered that the current waves which have travelled along one conductor will return along the other and in so doing will again produce a standing wave pattern with results as before. In this condition, the receiving end becomes a point of low impedance, a

place where V is small and I is big. The places of the nodes of current and voltage of Fig. 32 (*b*) are interchanged, and the new state of affairs is shown in Fig. 32 (*c*). This may be clearer if reference is made to Fig. 12; the dotted travelling waves of that case represent the current waves of this case, and show that a current anti-node will occur at the short circuited point. Similarly, Fig. 13 represents the voltage waves of this case.

In either case, the impedance presented by such a resonant feeder line will vary with the relative distance of the source from the receiving end. Use may sometimes be made of this fact in matching transmission lines to an aerial.

The use of **tuned feeders**, or transmission lines in which standing waves are present, is open to objection on grounds of efficiency. At the anti-nodes of potential large dielectric losses occur, and at the anti-nodes of current large ohmic losses occur, and it is generally considered desirable to suppress standing waves in feeders where high overall efficiency is desired. It should now be evident that anything which will prevent the reflection of energy will also prevent the production of standing waves. The best way to prevent the reflection of energy is to use it, by providing a suitable terminal impedance to dissipate energy at the rate at which it is supplied. If no energy is reflected, standing waves will not appear; no resonance effects will be observed. The feeder line will only have travelling waves in it, and will behave as if it were of infinite length. To make a feeder line equivalent to a line of infinite length, it is clear that at the receiving end one must add an impedance equal to that presented by an infinitely long line.

Now, it can be shown that an infinitely long line, possessing distributed capacity and inductance, has the property of a resistance, the value of which is given by

$$R = \sqrt{\frac{L}{C}}.$$

Energy could be continuously fed into such a line with a ratio of voltage to current equal to R, and for all practical purposes the line would act exactly as a resistance, the line continuously absorbing the energy put into it and passing it continuously forward to infinity. Allowing for ohmic and other losses this is represented in Fig. 30 (*d*). The resistance which would be presented by a feeder line if it were of infinite length is called the "**characteristic impedance**"; it is also called the "**iterative**" or "**surge impedance**". The term "surge impedance" is one which has been borrowed from low frequency power engineering. Its origin lies in the H/F surges of power that can be occasioned in low frequency power lines by certain causes, principally lightning flashes. It was observed in paragraph 13 that under ordinary conditions in L/F transmission lines, resonances or standing waves are unlikely to occur.

H/F resonance is, however, possible in those lines, and H/F surges can occur and can be very destructive unless steps are taken to prevent their possibility.

It can be shown that when the impedance at the receiving end has any value other than the surge impedance, partial absorption and partial reflection will occur.

The whole subject of transmission lines with distributed constants is a difficult one to handle mathematically, although there are various approximate treatments, two of which may be outlined as follows :—

(*a*) With reference to Fig. 32 (*b*), each half-wavelength may be considered equivalent to a tuned acceptor circuit of inductance L and capacity C which gives the same current and voltage ratio as exists on the feeder at the frequency concerned. The peak value of the current is therefore given by

$$I = \frac{V}{\omega L}$$

and since the circuit is tuned we have

$$\omega = \frac{1}{\sqrt{LC}}.$$

By substitution we have

$$I = V\sqrt{\frac{C}{L}} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

Now assume that the standing wave starts travelling along the feeder towards the receiving end P. It will carry its stored energy with it. If we assume that the receiving end, Fig. 32 (*a*), is closed by the surge impedance R, the peak value of the current flowing through it will then be given by

$$i = \frac{V}{R} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (2)$$

If there is no reflection, $i = I$ ...... and from equation (1)

$$\frac{V}{R} = V\sqrt{\frac{C}{L}}$$

that is $R = \sqrt{\frac{L}{C}}$.

(*b*) As we have already observed, suppression of standing waves will be achieved, if the feeder can be made equivalent to a non-resonant circuit.

An interesting result which shows the possibility of making a circuit non-resonant, is provided by the circuit of Fig. 33.

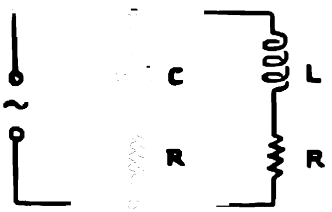


FIG. 33.

The total impedance is given by—

$$\frac{1}{Z} = \frac{1}{R - \frac{j}{\omega C}} + \frac{1}{R + j\omega L} = \frac{R_r + \frac{j}{\omega C}}{R^2 + \frac{1}{\omega^2 C^2}} + \frac{R - j\omega L}{R^2 + \omega^2 L^2}$$

$$= \frac{R}{R^2 + \frac{1}{\omega^2 C^2}} + \frac{R}{R^2 + \omega^2 L^2} + j\left(\frac{\frac{1}{\omega C}}{R^2 + \frac{1}{\omega^2 C^2}} - \frac{\omega L}{R^2 + \omega^2 L^2}\right).$$

The condition for resonance is given by equating the "j" parts to zero.

$$\text{i.e.} \quad \frac{\frac{1}{\omega C}}{R^2 + \frac{1}{\omega^2 C^2}} = \frac{\omega L}{R^2 + \omega^2 L^2}$$

$$\text{or} \quad R^2 + \omega^2 L^2 = \omega^2 LC\left(R^2 + \frac{1}{\omega^2 C^2}\right) = \omega^2 LCR^2 + \frac{L}{C}$$

$$\text{or} \quad \omega^2 (L^2 - LCR^2) = \frac{L}{C} - R^2$$

$$\therefore \quad \omega^2 = \frac{1}{C}(L - CR^2) \cdot \frac{1}{L(L - CR^2)} = \frac{1}{LC} \cdot \frac{L - CR^2}{L - CR^2}$$

Hence if $L = CR^2$ or $R = \sqrt{\frac{L}{C}}$, the resonant frequency may have any value between zero and infinity; in fact, the circuit is **non-resonant.**

It may be shown, moreover, that the impedance of the above circuit when the resistances be of value given by $R = \sqrt{\frac{L}{C}}$ is independent of the frequency of the supply.

The Total Impedance is given by :—

$$\frac{1}{Z} = \frac{1}{R - \frac{j}{\omega C}} + \frac{1}{R + j\omega L} = \frac{1}{R + \frac{1}{j\omega C}} + \frac{1}{R + j\omega L}$$

But $\sqrt{\frac{L}{C}} = R \quad \therefore \frac{L}{C} = R^2$ and $\omega L = \omega CR^2$

$$\text{Hence } \frac{1}{Z} = \frac{j\omega C}{1 + j\omega CR} + \frac{1}{R + j\omega CR^2} = \frac{j\omega CR + 1}{R + j\omega CR^2} = \frac{1}{R}\left(\frac{1 + j\omega CR}{1 + j\omega CR}\right) = \frac{1}{R}$$

$$\therefore Z = R = \sqrt{\frac{L}{C}}$$

The Impedance at all frequencies is thus equal to R, provided that the two resistances are each made of value equal to $\sqrt{\frac{L}{C}}$.

When L and C represent the uniformly distributed inductance and capacity of a line, their ratio will be unaffected if their values are referred to unit length.

Actual values of L and C will depend on the type of feeder in use.

37. **Types of Feeder Lines.**—Essentially there are two varieties :—

(*a*) The parallel wire or twin feeder, and

(*b*) The concentric tube feeder.

The TWIN FEEDER consists of two wires spaced as closely together as the voltage between them will allow, and mounted on poles in a plane parallel to the ground and 7 ft. or more above it. The spacers between the wires must be sufficiently close to maintain a constant wire distance, and the span between the poles must be short enough to prevent the system from swaying ; if this is not done, the capacity between each wire and earth will vary, and cause the wave frequency and output to vary in the case of self-oscillatory circuits but the output only in the case of master controlled transmitters. Fig. 34 shows diagrammatically a twin feeder making connection between the transmitter and the aerial coupling circuit. An advantage of the twin wire feeder is that at any point along it the currents in the two wires are equal and opposite in sign, and, since the wires are close together, the radiation from the one is

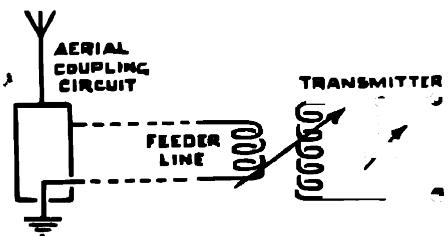


FIG. 34.

cancelled by that from the other. The production of a non-radiating feeder system is a matter of considerable importance in both transmitting and receiving aerial systems. Quite a small amount of radiation from the feeder is sufficient to spoil the directional properties of the aerial.

It may be asked why the coupling coil is not directly earthed in order to save the length of wire between the coil and the aerial coupling circuit. If this were done, the single feeder wire would radiate energy like an aerial and very little would go into the aerial proper. A single feeder line with an earth return can only be used in the non-resonant state; it is not easy to achieve the necessary impedance matching to effect this condition, and, in practice, twin feeder lines balanced to earth are to be preferred.

Consider the case of a return circuit, or feeder line, consisting of two parallel wires of radius "*a*", each of length "*l*" and spaced "*d*" apart, the currents in the two wires being opposite in sign. When air is the dielectric and when the ratio d/l is small it can be shown—

$$L_1 = 4l \log \frac{d}{a} \quad \text{... (where } L_1 = \text{total inductance).}$$

$$\therefore \qquad L = \frac{L_1}{l} = 4 \log \frac{d}{a} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

where L = inductance per unit length).

Using logs to base 10, mics., and cms., this becomes

$$L = 9{\cdot}21 \times 10^{-3} \log_{10} \frac{d}{a} \quad \ldots\ldots \quad \text{in } \frac{\mu H.}{cm.}$$

Similarly for capacity we can show—

$$C = \frac{C_1}{l} = \frac{1}{4 \log \frac{d}{a}} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (2)$$

and using logs to base 10, μF's and cms., this becomes

$$C = \frac{10^{-4}}{828 \log_{10} \frac{d}{a}} \quad \ldots\ldots \quad \text{in } \frac{\mu F.}{cm.}$$

Expressing (1) in E.S.U's we have

$$L = \left(4 \log \frac{d}{a}\right) \frac{1}{9 \times 10^{20}}$$

$$\frac{L}{C} = \left(4 \log \frac{d}{a}\right)^2 \frac{1}{9 \times 10^{20}}$$

$$\sqrt{\frac{L}{C}} = \left(4 \log \frac{d}{a}\right) \frac{1}{3 \times 10^{10}} \quad \ldots\ldots\ldots \quad \text{in E.S.U's.}$$

and in ohms this becomes

$$\sqrt{\frac{L}{C}} = \left(4 \log \frac{d}{a}\right) \frac{9 \times 10^{11}}{3 \times 10^{10}} = 120 \log \frac{d}{a}$$

and using logs to base 10, this becomes

$$\sqrt{\frac{L}{C}} = 276 \log_{10} \frac{d}{a} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (3)$$

(a well-known formula).

Formula (3) gives values for the surge impedance between 400 and 800 ohms over a wide range of practical values of the ratio d/a. In general, it is convenient to use a transmission line of such dimensions as to give a surge impedance of 600 ohms, although for special purposes some higher or lower figure may be an advantage.

It is important to note that formula (3) cannot be used to give the value of the surge impedance of twin rubber covered flexible wire, for in that case the dielectric is no longer air. The characteristic impedance of flexible wire of the type used for lighting is usually somewhere between 150 and 200 ohms, the figure for somewhat heavier flexible wire being in the region of 100 ohms.

At Horsea Island W/T Station the parallel wire feeder system consists of two wires 0·192 ins. in diameter, spaced 6 ins. apart, and carried at a height of 13 ft. from the ground by poles spaced approximately 70 ft. apart. It can be verified that this gives a value of surge impedance in the neighbourhood of 495 ohms. At the B.B.C. station at Droitwich, the feeder line from the transmitter working on 200 kc./s. is also of the parallel wire type. It is supported on poles 13 ft. high and has a characteristic impedance of 560 ohms.

Parallel wire feeders are simple to erect, but are inclined to suffer seriously from radiation unless the lines are very carefully terminated. At the higher frequencies this may often be a matter of some difficulty. Twin feeders also suffer from the fact that they are exposed to the weather; certain of the feeder losses become variable ones, and the surge impedance may alter. It is usual to cross the feeders at regular intervals to reduce mutual action between sections.

The CONCENTRIC TUBE FEEDER, in which one conductor completely encloses the other, represents an approach to the ideal from the point of view of electrical isolation. Moreover, the circular symmetry of the line makes it much more susceptible to exact mathematical analysis than can be the case with open wire lines. In practice the outer tube of the feeder is earthed at numerous points along its length; in receiving systems where it is specially important to avoid picking up static and other electrical disturbances, it might even be advantageous to bury the feeder in the ground. Concentric tube feeders are especially applicable to aerial systems working on a fixed frequency. Their prime cost is, however, much greater than that of a twin wire system, largely owing to the greater amount of copper that has to be used. Concentric tube lines may be made weather proof and the insulators inside the outer earthed tube are protected; losses due to this cause are likely to be smaller and more constant than with twin wire feeders.

In the case of concentric tube lines, in which " a " is the outer radius of the inner conductor and " b " is the inner radius of the outer conductor, the expression for the **inductance per unit length** was first derived by Lord Rayleigh, assuming the two tubes to be of negligible thickness. It is given by—

$$L = 2 \log \frac{b}{a} \quad \ldots\ldots\ldots \quad \text{in E.M.U's.} \quad \ldots\ldots\ldots \quad (4)$$

The formula for the **capacity per unit length** of such a system is better known, and more easily proved. With air as dielectric, it is given by—

$$C = \frac{1}{2 \log \frac{b}{a}} \quad \text{in E.S.U's.} \quad \ldots\ldots\ldots\ldots\ldots \quad (5)$$

Expressing (4) in E.S.U's

$$L = \left(2 \log \frac{b}{a}\right) \frac{1}{9 \times 10^{20}}$$

$$\frac{L}{C} = \left(2 \log \frac{b}{a}\right)^2 \frac{1}{9 \times 10^{20}}$$

$$\sqrt{\frac{L}{C}} = \left(2 \log \frac{b}{a}\right) \frac{1}{3 \times 10^{10}} \quad \ldots\ldots\ldots \quad \text{in E.S.U's.}$$

and in ohms this becomes

$$\sqrt{\frac{L}{C}} = \left(2 \log \frac{b}{a}\right) \frac{9 \times 10^{11}}{3 \times 10^{10}} = 60 \log \frac{b}{a}$$

and using logs to base 10 this becomes

$$\sqrt{\frac{L}{C}} = 138 \log_{10} \frac{b}{a} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (6)$$

(a well-known formula).

Formula (6) usually gives values of the surge impedance in the neighbourhood of 70 ohms over a range of practical values of the ratio b/a. In one of the feeder systems of the Marconi Company, the outer earthed tube is $3\frac{1}{2}$ ins. in diameter, and the inner live conductor is a tube $\frac{7}{8}$ in. in diameter and insulated from the outer sheath by spider porcelain insulators at various intervals. The whole system is supported on stakes at a height of about 1 ft. from the ground. It has been calculated that to minimise copper losses there is an optimum ratio of diameters of about 4 : 1. The B.B.C. have installed concentric feeders at the Daventry station for use in connection with the Empire Beam Service, and more recently at Lisburn in connection with the North Ireland Regional Station.

Co-axial cable is a comparatively new commercial product which has been developed in connection with television. The central conductor is separated from the outer co-axial one by means of an air dielectric, and in one specification the cable is shown to possess a characteristic impedance of about 67 ohms, a capacity of 0·095 microfarads per mile, and an inductance of 0·415 millihenries per mile. In that case, the central conductor is held in position by kinking the wire successively twice in planes at right angles, the sets of kinks being equally spaced throughout the length of the cable. It appears possible that this type of cable may find an increasing number of uses ; it could be used very conveniently as a transmission line joining a high frequency aerial system to its receiver. It is usually called " co-axial " cable to distinguish it from **concentric cable,** which has long been in use in low frequency A.C. engineering.

The relative superiority of a concentric feeder line is very marked in the case of reception.

38. **Matched Systems.**—With reference to paragraph 35, it is now possible to consider the design of complete systems to achieve the necessary impedance matching between the unit delivering power at one end of the line, and the power absorbing resistance (the aerial) at the other end. It is stressed that the object of insistence on this impedance matching is the necessity for the efficient transference of energy between the transmitter and the aerial. Mechanically, the process is analogous to the propulsion of a vehicle along a road ; with reference to the power of the engine, it is necessary to choose the gear appropriate to the gradient, *i.e.*, the rate at which work has to be done.

Impedance matching between aerial and transmitter may be achieved by using—

(*a*) **untuned feeder lines** ; this implies terminating each end of the line by its surge impedance, or

(*b*) **tuned feeder lines** ; in this case standing waves will be present on them and they are usually known as Lecher wires (Lecher—a German scientist who produced standing waves in wires in 1890).

At H/F it will be more efficient to employ untuned feeders, although for short distances tuned feeders may be permissible. Even at L/F, when the transmitter and aerial are relatively adjacent, it is sometimes desirable to use an untuned transmission line ; this is the case at the B.B.C. station at Droitwich, where a transmission line of length $0 \cdot 15\lambda$ is in use, in connection with the transmitter working on 200 kc./s.

It is proposed to consider a number of simple matched systems :—

(i) $\lambda/4$ AND $\lambda/2$ AERIALS WITH UNTUNED FEEDERS.—Taking the case of a line with a surge impedance of 600 ohms, the aerial must present to the line a resistance of the same value. In general, this value is not obtainable by direct connection to the aerial. It was seen in paragraph 11 that a quarter wave aerial has a radiation resistance of the order of 40 ohms, and at its current anti-node it behaves as an impedance of about that value. It was also shown that a half wave aerial has a radiation resistance of about 80 ohms, and at its current anti-node it also behaves as an impedance of about that value ; at its current node it behaves as an impedance of much higher value, intermediate values being observed on intermediate points. In either case the line impedance is of a totally different order.

It may occasionally be possible to terminate a transmission line with the aerial itself ; for example, a concentric feeder, or other line with surge impedance of about 80 ohms, could probably be satisfactorily terminated by the middle of a half wave aerial, without the help of any matching devices. Among amateur transmitters using low power, it is quite common practice to feed the centre of a half wave aerial by means of twisted flexible wire of the ordinary variety used for lighting ; quite good matching is obtained in this way.

The linking up of a power unit to a transmission line may be done in several ways, and can be done by means of a transformer coupling. From simple transformer theory it can be appreciated that a 600 ohm line may be linked to a 5,400 ohm impedance, such as that often found in loud-speaker networks, by means of an *iron cored* step-up transformer having a ratio of 3 : 1. The calculation for T, the transformer ratio, may be set out as follows—

$$\frac{5400}{T^2} = 600.$$

The provision of high impedance matching lines is inexpedient, on account of their very excessive spacing.

Fig. 35 represents diagrammatically a matched system, in which approximate matching is obtained at each end of the feeder line. The *air-cored* step-down transformer at the aerial end makes the feeder line present an impedance of roughly 40 ohms to the aerial circuit. At the transmitter end, the step-up transformer presents the requisite matching load to the **tank circuit.** Fig. 36 presents another circuit achieving the same result, by using auto-transformer couplings. For simplicity it may be assumed that Figs. 35 and 36 represent a C.W. transmitter working on a fixed frequency.

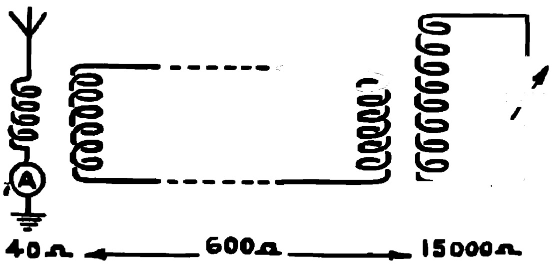


FIG. 35.

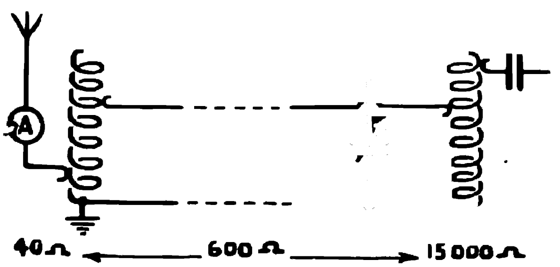


FIG. 36.

(ii) THE TRANSDUCER (TRANSMISSION LINE IMPEDANCE REDUCER).—The problem of matching a telephony aerial is a peculiar one, in view of the fact that its impedance varies over the side band frequency range. The ideal telephony aerial for broadcast purposes would be one in which the resistance or reactance did not change appreciably over the range of frequencies represented by the carrier frequency $\pm$ about 8 kc./s. Although it might be possible to design such an aerial, it was not found to be sound policy to do so in the case of the B.B.C. station at Droitwich. In order to produce an aerial system with a reasonably flat frequency

response curve over the range of the side band spectrum, it was necessary to proportion the aerial constants so that the ratio reactance/resistance should not be too big. A suitable value was obtained by using relatively high masts to give a high value of the radiation resistance Serious attenuation of the side band frequencies, corresponding to the higher modulating frequencies, would have resulted if the aerial had been coupled directly to the output circuits of the transmitter. It was in order to eliminate this difficulty that a relatively complicated impedance matching system was designed, which has since become commonly known as a "transducer." Fig. 37 represents

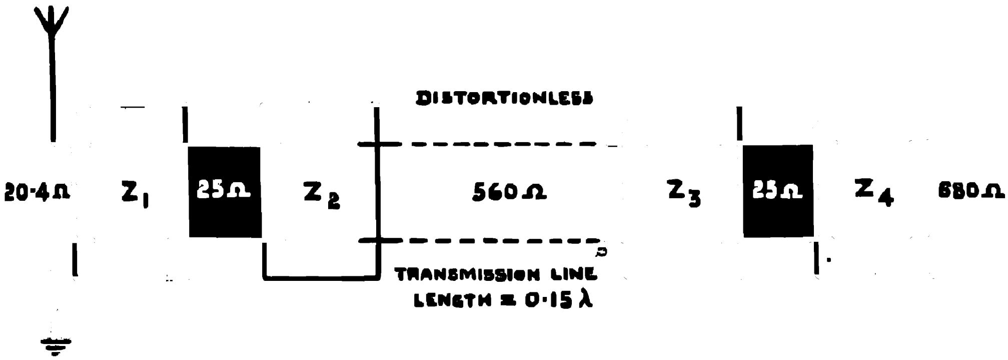


FIG. 37.

a very much simplified schematic diagram of the transducer, and it should be remembered that the various parts of the network have functions other than mere impedance matching. Their functions may be simply stated as follows :—

Z_4......(*a*) Presents the requisite impedance (*i.e.*, 680 ohms) to the tank circuit of the transmitter on the carrier frequency, the impedance-frequency characteristic being symmetrical for 8 kc./s. on either side of 200 kc./s., the carrier frequency.

(*b*) Converts 680 ohms to 25 ohms.

Z_3......(*a*) A reactance compensating network that makes it possible to match up to a distortionless transmission line (*i.e.*, one in which the transfer parameter is independent of the frequency).

(*b*) Converts 25 ohms to 560 ohms, the surge impedance of the transmission line.

(*c*) A low pass filter to assist in R/F harmonic suppression.

Z_2......(*a*) Converts 560 ohms to 25 ohms.

(*b*) A reactance compensating network that allows for the change in reactance of the aerial over the transmitted frequency band, and so prevents attenuation of the side band frequencies corresponding to the higher modulating frequencies.

Z_1......(*a*) Converts 25 ohms to 20·4 ohms, which is the aerial resistance at its natural frequency of 200 kc./s.

(iii) THE CENTRE FED λ/2 AERIAL, WITH AN UNTUNED FEEDER: STUB MATCHING.—H/F aerials are often of the λ/2 type, and quite frequently it is desirable to arrange that the radiating portion shall be horizontal. To avoid radiation from the vertical leads to the centre, it is important to have a carefully designed feeder system. Fig. 38 represents a spaced wire vertical feeder line, supplying energy to a horizontal dipole aerial. At the

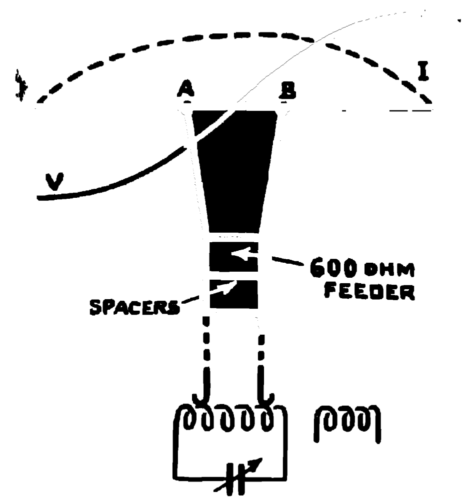


FIG. 38.

top, the feeder line is tapped into the aerial so that the aerial impedance at the tapping point is that required for matching the feeder. At the lower end, the matching of the feeder to the tank circuit is accomplished by adjusting the taps on the inductance. These two taps should be moved together up and down the coil in order to equalise the currents in the two feeder wires.

The principle of matching a dipole aerial to a transmission line by moving the tapping points A and B is an important one, and its application may be seen in other forms. In one form, the $\lambda/2$ dipole is split in the middle and a portion of each part—the stub end—is folded back in a plane at right angles, so as to form a part of the transmission line. The folded back portions will still carry standing waves, but their radiating effect will be mutually cancelled; so far as impedance is concerned they behave as does the aerial itself. There are, therefore, along these wires points of high and points of low impedance, and it is only necessary to connect up (say) a transmission line of 600 ohms to points on the aerial continuation elements at which the impedance is also 600 ohms; these elements may be connected together at their ends, or they may be left open, the matching points being differently situated in the two cases. This method of terminating an aerial is called "stub matching."

(iv) THE HORIZONTAL $\lambda/2$ "ZEPPELIN" AERIAL, WITH TUNED FEEDER.—Another method of feeding energy to a horizontal $\lambda/2$ aerial is shown diagrammatically in Fig. 39. In this case a tuned feeder is used, but if the currents in each of the vertical leads is adjusted to equality, the radiation from them should be practically nil. The radiating portion AB may be any number of half-waves in length, and the length of the feeder lines should be adjusted to be an odd number of quarter wavelengths, if the whole circuit is tuned as an acceptor circuit—as shown in Fig. 39. If it is desired to use **parallel tuning,** as in Fig. 40 (*b*), the length of the feeder lines should be an even number of quarter waves in length.

This type of radiator is usually known as a "Zeppelin" aerial, and is very popular among amateur transmitters.

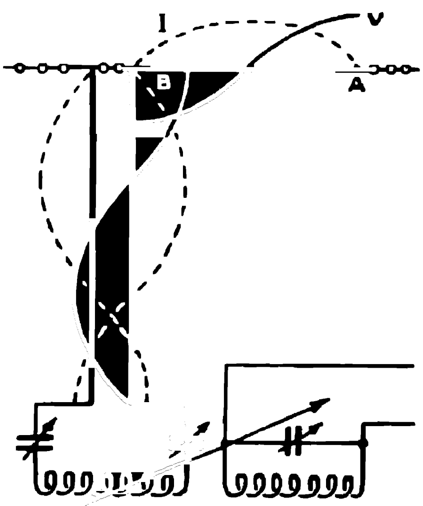


FIG. 39.

(v) THE HORIZONTAL CENTRE FED $\lambda/2$ AERIAL, WITH A TUNED FEEDER.—Figs. 40 (*a*) and (*b*) represent a split $\lambda/2$ aerial being energised in two different ways. The feeders are tuned, but the radiation from the vertical leads is mutually cancelled.

Fig. 40 (*a*) shows the voltage and current distribution when the horizontal radiating portion is a half wave in length. The figure shows that the circuit is tuned as an acceptor circuit, since there is a voltage node at the two condensers. In this case, therefore, the length of the feeder lines must be adjusted to be an even number of quarter waves in length.

Fig. 40 (*b*) shows the state of affairs when the top portion is a whole wave in length, the aerial system in this case being arranged to tune as a rejector circuit. The length of the feeder lines must therefore be adjusted to be an even number of quarter waves in length, in order to produce a voltage anti-node across the condenser.

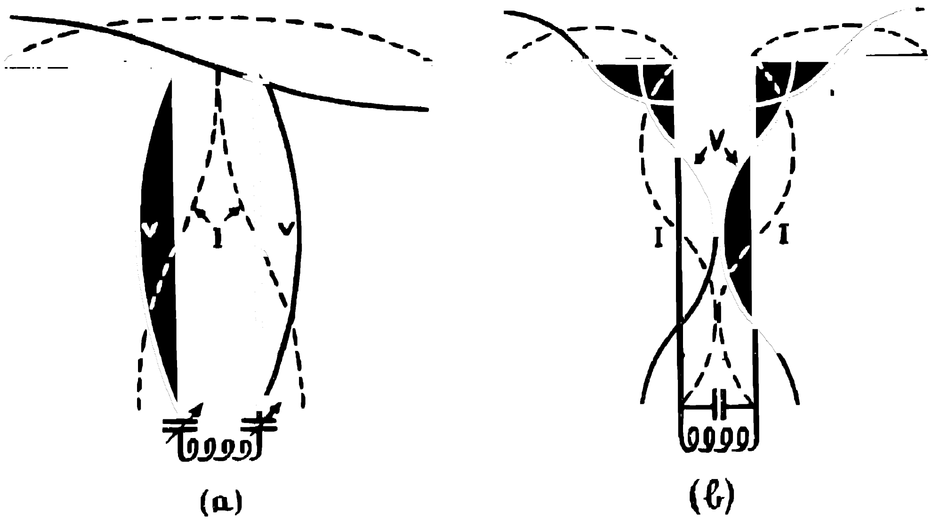


FIG. 40.

(vi) THE UNTUNED FEEDER AND QUARTER WAVE MATCHING LINE.—When tuned transmission lines have perfectly distributed constants, the impedance will be the same at successive current anti-nodes and the state of affairs will be that represented in Fig. 32. If the line is a half wave in length, the impedance at each end will be the same, and we may write—

$$Z_s = Z_r \quad \ldots\ldots\ldots\ldots \quad (1)$$

(where Z_s and Z_r are the sending and receiving end impedances).

Equation (1) means that a line one half wave in length behaves as a 1 : 1 transformer. The ratio between the impedance at each end of a line a quarter wave in length depends upon $\sqrt{\frac{L}{C}}$, and from transmission line theory beyond the scope of this book, we have the relation

$$Z_s Z_r = Z_0^2 \quad \ldots\ldots\ldots\ldots \quad (2)$$

$$\left(\text{where } Z_0 = \sqrt{\frac{L}{C}}\right).$$

This relation is obviously true in the special case when no reflection occurs and

$$Z_s = Z_r = Z_0 = \sqrt{\frac{L}{C}}.$$

Equation (2) means that the tuning properties of a line one quarter wave in length, enable it to be used as a step-up or step-down transformer, the ratio for a given case depending on the surge impedance. This has led to the development of " quarter wave matching lines," these lines being

so constructed as to have a characteristic impedance equal to the square root of the product of the impedances to be linked. Thus, if a 600 ohm transmission line is to be linked up to a 3340 ohm aerial, the matching line would have a characteristic impedance of $\sqrt{3340 \times 600} = 1414$ ohms. A $\lambda/4$ impedance matching section may occur at the end of a feeder line, and may consist of an open feeder the spacing of which can be varied. Fig. 41 represents the use of this system with a centre-fed dipole aerial, with standing waves as indicated. The process of matching the impedance consists in varying the spacing until there is no reflection in the transmission line.

Reflection in transmission lines may be detected by inserting three or more R/F ammeters at points about $\lambda/6$ apart; if the currents indicated at such points are equal, it may be concluded that standing waves are not present, and that the line is terminated with a load equal to its surge impedance. This is perfectly general method, which is applicable to any transmission line.

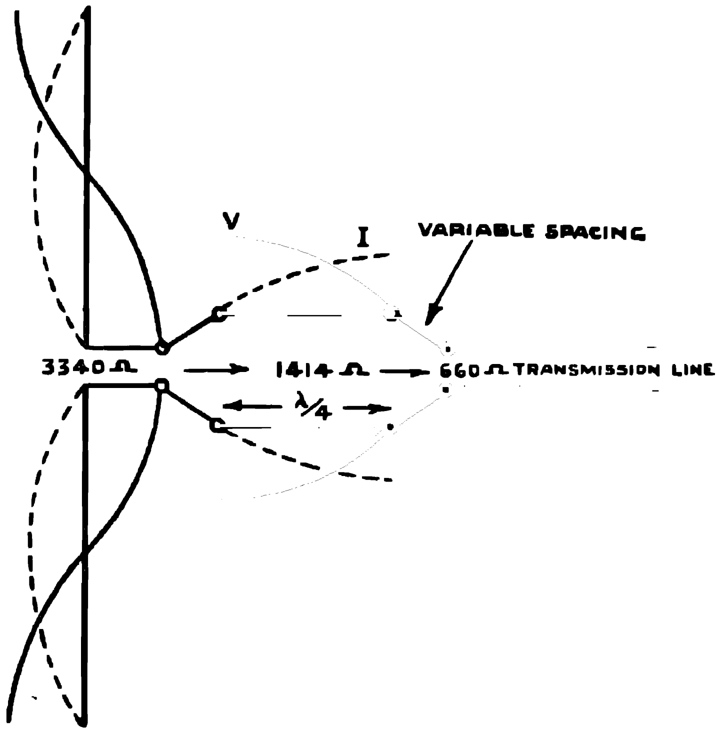


FIG. 41.

(vii) THE QUADRUPOLE.—Any two dipoles connected in parallel across a feeder line may be said to constitute a "quadrupole," and Fig. 42 (*a*) shows one consisting of two crossed dipoles. If each dipole presents an impedance R to the transmission line, the two in parallel will present an effective impedance of R/2, as shown in Fig. 42 (*b*), and the transmission line must be designed accordingly. Fig. 42 (*c*) represents a horizontal quadrupole suspended from triatics; the system is supplied by a vertical feeder which is terminated and energised at its lower end by means of a push-pull circuit. If the horizontal radiating portions are each a half wave in length, it is possible that the impedance of each separately might be of the order of 2,000 ohms, so that jointly they would present an impedance of about 1,000 ohms to the vertical feeder line. If the dipoles each have the same physical dimensions, it is clear that the arrangement would be suitable for use at one frequency only. If, however, each dipole has a different natural frequency, it has been stated that the arrangement will give reasonably flat tuning over a band of frequencies, the centre of which lies roughly midway between the two fundamental frequencies.

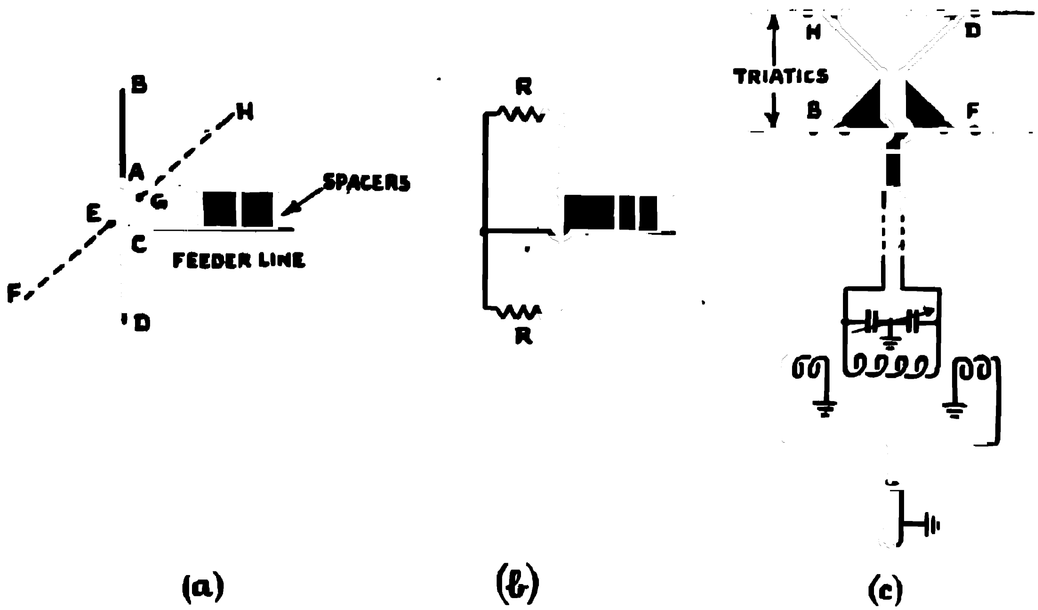


FIG. 42.

39. **Polar Diagrams.**—In paragraph 12, it was observed that, from the point of view of communication range, the radiation efficiency provided no criterion of performance ; each aerial is designed for its special function, and the *distribution* of the radiated power is more important than its total numerical value. In the special case of M/F broadcast work the chief requirement is that equal power shall be radiated in all directions in the horizontal plane ; any energy which is radiated in upward directions not only produces inefficiency but is also responsible for the fading which occurs at the limits of the **service area,** due to the inter-action of the direct and indirectly radiated waves. For broadcasting services, where communication with aircraft is not desired, and where the object is to " heat the wires of the receiving station and not the universe " (Stuart Ballantine), it is better to consider the **power efficiency to produce a given field strength on the ground** rather than the radiation efficiency ; in the extreme case, the radiation efficiency might be high but all of the energy might be radiated vertically.

The B.B.C. high-frequency Empire Beam services depend for their action upon reflection from the ionosphere, and provide another special case in which it has been shown that correct directivity in the vertical plane is at least as important as that in the horizontal plane. This implies that the radiation from the aerial can be directed in the same way as the path of a projectile from a gun.

In three dimensional space, the behaviour of any aerial or aerial system may be expressed by polar diagrams. These diagrams may be *horizontal* or *vertical* ones, and represent the relative values of either field strength or power radiated at various angles in the horizontal or vertical planes, the pole being taken at the foot of the aerial or the centre of the aerial system. In many simple cases polar equations can be calculated from which the curves can be plotted ; the mathematical analysis is troublesome and will not be given here. The length of the line from pole to curve at any angle is a measure of the *relative* field strength at that angle.

40. **Polar Diagrams for Vertical Aerials.**—The work of paragraph 8 referred to a simp radiator in a free space in which the current was uniform along its length. In that case, the field strength was shown to vary as the cosine of the angle of elevation of the ray from the horizontal, giving a vertical polar diagram having the form of a figure-of-eight. In the case of an earthed aerial, the length of which is short in comparison with a quarter wavelength, the form of the vertical polar diagram is approximately semi-circular (a cosine diagram), if one assumes that the earth is

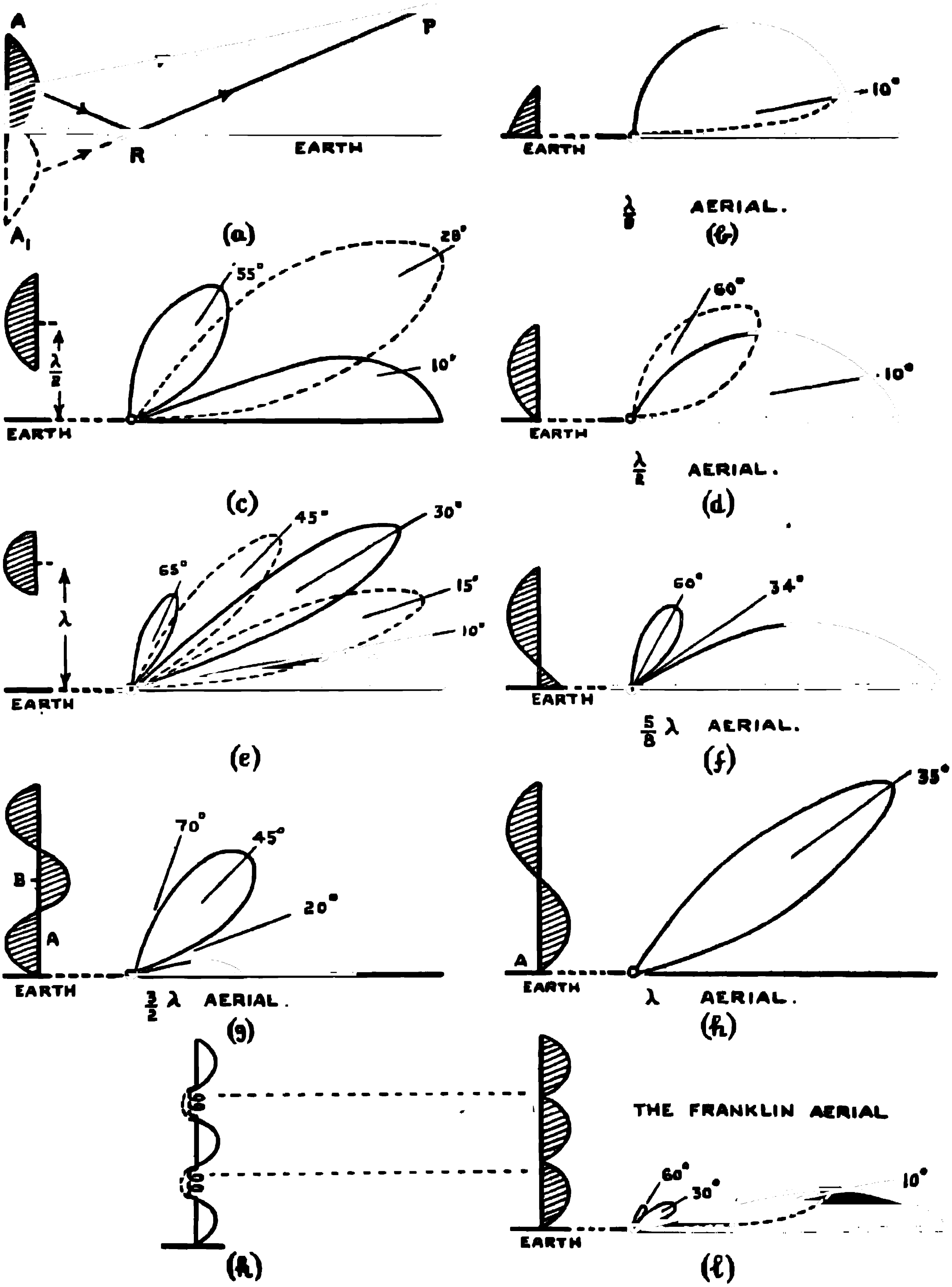

A
P
R
EARTH
A_1
(a)
10°
EARTH
$\frac{\lambda}{8}$ AERIAL.
(b)
28°
55°
$\frac{\lambda}{2}$
10°
EARTH
(c)
60°
10°
EARTH
$\frac{\lambda}{2}$ AERIAL.
(d)
45°
30°
65°
15°
10°
λ
EARTH
(e)
60°
34°
EARTH
$\frac{5}{8}$ λ AERIAL.
(f)
70°
45°
B
20°
A
EARTH
$\frac{3}{2}$ λ AERIAL.
(g)
35°
A
EARTH
λ AERIAL.
(h)
THE FRANKLIN AERIAL
60°
30°
10°
EARTH
(k)
(l)

FIG. 43.

a perfect conductor. As the length of the aerial becomes commensurate in value with the wavelength in use, the current variations along it become more pronounced, phase differences appear, and interference effects between components of the radiation from various parts of the aerial may produce *reinforcement* of the radiation in certain directions and *cancellation* in others.

Fig. 43 (*a*) represents an earthed aerial of length slightly greater than a quarter wave. At some distant point P, energy may arrive via the direct path AP, or by the indirect path ARP. Provided that the energy arriving at P via each path is similarly polarised, cancellation will occur when the difference in path length is equal to an odd number of half wavelengths, for in that case waves will arrive 180° out of phase with each other. Using the conception of "**electrical images**" it has been shown that an effect at P may be assessed by adding together the radiation due to the direct ray with that due to the image of the aerial in the earth.

The earth acts as an imperfect mirror to an electro-magnetic wave striking it. Provided the reflecting area is damp soil or the sea, over 80 per cent. of the energy in the wave will be reflected. If it is dry sand or heavily wooded, the percentage will be much less. The hilliness of the ground will also affect the results, the earth being a much better reflector for frequencies corresponding to long waves than those corresponding to short ones. At H/F a high hill becomes comparable in size with the wavelength in use, and produces a relatively greater effect. The angle of incidence is equal to the angle of reflection, and for some practical purposes we can consider the reflected wave as being in phase with the incident wave.

In the basic theory of reflection of waves that are electro-magnetic in character, it has been proved that there is a certain critical angle of incidence for vertically polarised waves, after which a change of phase of 180° occurs in the reflected wave. This critical angle is usually known as **Brewster's angle** (Sir David Brewster, 1811); it varies with the wave frequency and the earth constants. Measured in the way shown in Fig. 43, it becomes greater as the frequency increases and the earth becomes a less perfect conductor. Brewster's angle is also sometimes known as the "**angle of polarisation.**" This has reference to the fact that if non-polarised electromagnetic waves —such as ordinary light waves—are incident on a reflecting medium at this critical angle, only the horizontally polarised components are reflected.

For aerials where the vertical height is small in comparison with a quarter wavelength, there is not sufficient difference in phase between the current in the real aerial and its image to cause appreciable cancellation of radiation in upward directions. The polar diagram is approximately semi-circular, as shown in Fig. 43 (*b*). When the height of the aerial is increased to a half wavelength, it is clear that the point of maximum current in the real aerial is distant by a half wavelength from the similar point in the image aerial; in this case it becomes possible for the direct and reflected rays at some distant and elevated point P to have a difference in path length of an odd number of half wavelengths. Considerable cancellation of high angle radiation may therefore take place, and for the same aerial input power the relative form of the vertical polar diagram is shown in Fig. 43 (*d*). It will be noted that it is very much flattened out, representing a concentration of the radiated energy in more horizontal directions.

If the height of a vertical aerial is made greater than a half wavelength, the diagram is still further flattened but a secondary lobe of radiation appears at an angle of about 60° from the horizontal—Fig. 43 (*f*). It has been shown theoretically that for a maximum power efficiency to produce a given field strength on the ground, the best height of the aerial should be $0 \cdot 625\lambda$. In practice, a height of the order of $0 \cdot 58\lambda$ is used, and this is the principle of design of the new "**anti-fading**" aerials for use in M/F broadcast work. These diagrams are calculated on the assumption that the earth has perfect conductivity. This is an assumption which at high frequencies produces results which are very wide of the mark. In all cases in Fig. 43 the full line curves show the variation in field strength at different vertical angles assuming a *perfectly* conducting earth. The dotted curves attribute *negligible* conductivity to the earth's surface; they were developed analytically in 1927 by T. L. Eckersley. They are difficult to calculate and the earth's "constants" are very variable. The more nearly perfect the conductivity of the earth becomes, the more nearly the dotted curve approaches the full line curve. Fig. 43 (*b*) explains, incidentally, the well known effect that high frequency reception improves considerably as the receiver is taken up a hill having a

downward slope towards the transmitter. In practice the effect of an imperfectly reflecting earth also means that there is no angle of zero radiation, but merely a minimum.

Taking an average value for the earth's constants, at H/F the polarising angle is of the order of 10°. Eckersley showed that for angles above this, the polar diagram is that which is appropriate to a perfectly conducting earth, namely, the full line curves of Fig. 43, and for angles below this it is that which is appropriate to an earth of negligible conductivity—the dotted curves of Fig. 43.

The vertical polar diagrams for a vertical half wavelength aerial whose centre is at various heights above the earth's surface are shown in Fig. 43 (*c*) and (*e*). It will be seen that the effect of raising the aerial is to concentrate the radiated energy into " **beams** " at definite angles, and that these beams increase in number and become narrower as the height of the aerial above the ground is increased. The dotted curves represent the results which are usually obtained in practice due to an imperfectly conducting earth.

Vertical polar diagrams for certain other special cases are sketched in Fig. 43 (*g*), (*h*) and (*l*). In 1922, experiments were made with a long vertical aerial, the height of which was equal to the wavelength in use. Fig. 43 (*h*) shows the vertical polar diagram in that case. Maximum energy is radiated at an angle of about 35° to the horizontal. It will be noted that there is no horizontal radiation ; evidently this must be so since the oscillations in AB and CD are in phase opposition and, therefore, produce equal and opposite effects at a distant point in a horizontal plane through A. This will remain the case even if the aerial height is made equal to several wavelengths, provided that the current distribution is a pure sine wave throughout the aerial. In practice, the bottom half wave is greater than the top one, and there will be some ground radiation.

Considering the case of Fig. 43 (*g*), the current in the half wave section B is in anti-phase to that in the two other half wavelengths ; the resultant effect of such an aerial is to produce a beam at some angle to the horizontal. The form of the vertical diagram may be materially altered by arranging to suppress the anti-phase radiation from the section B. An aerial of this nature was developed by the Marconi Company in which alternate half waves were suppressed by inserting a coil at the current anti-nodes, the natural frequency of the coil being equal to that of the aerial. It is usually known as the **Franklin aerial**, and is shown diagrammatically in Fig. 43 (*k*) and (*l*) ; the coils are called " **phasing coils** " and have distributed capacity and inductance, but since they have practically no capacity to earth, they do not produce a radiation field. The radiating sections of the aerial now all radiate in phase, and Fig. 43 (*l*) shows that the main effect is to produce a concentration of energy in a beam at a lower angle than in three half wavelength aerials in which no suppression is employed. For high frequency long distance communication the burden of experimental evidence seems to be in favour of such **low angle radiation.**

A further development of phasing in aerials is to arrange the aerial wire itself according to such a pattern that the desired directions of maximum radiation are obtained without the use of phasing coils.

The radiation from any of the vertical aerials mentioned above is approximately symmetrical in any horizontal plane ; in three dimensional space this is represented by a solid figure of revolution having a cross section which is shown in the various cases in Fig. 43. This means that the vertical polar diagrams hold for any horizontal plane. This ceases to be the case when horizontal aerials are used, the radiation then becoming directional in the horizontal as well as in the vertical plane. In each of the above cases, however, the horizontal polar diagram is a circle. For directional horizontal radiation, **aerial arrays** are generally used.

41. **Polar Diagrams for Horizontal Aerials.**—These were investigated by Levy in 1924, and the connections to various horizontal aerials were described in paragraph 38.

A most interesting case is that in which a half wavelength dipole is placed at a height of a quarter wavelength above the earth. The aerial and its image in the earth are in phase opposition, but since they are distant from each other by a half wavelength, the fields of the space waves will reinforce each other in the direction of the zenith and will cancel each other in the horizontal direction In oblique directions, the path differences of the different rays from a point on the aerial and its

image grows less and there is a gradual diminution of the field to a zero value in the horizontal direction. This is shown in Fig. 44 (*a*). Since the radiation is horizontally polarised, there is no critical angle at which the polar diagram changes, as it did in the cases represented in Fig. 43.

Fig. 44 (*b*), (*c*), and (*d*) illustrate the changes which take place in the vertical polar diagram as the height of the horizontal doublet is increased.

Comparing the horizontal half wave doublet with a vertical one, it will be noted that in both cases the relative amount of horizontal radiation is increased by raising the aerial to a height. For high angle radiation a horizontal doublet with its centre a quarter wavelength above the earth is probably as good as any other aerial. Experimental work carried out in France and America on horizontal aerials tends to indicate that they possess very little advantage over vertical aerials working under the same conditions. With equal inputs, the strengths of the signals received are often about equal. There is, however, some evidence indicating that a degree of freedom from certain types of fading may be experienced if horizontal aerials are used. Moreover, in the case of

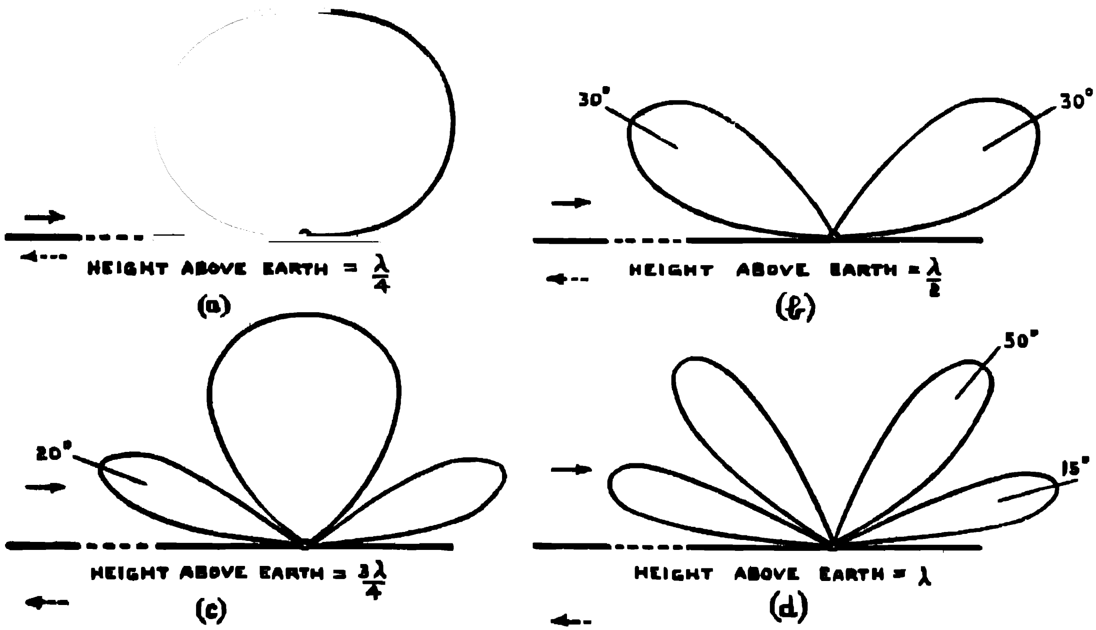


FIG. 44.

reception, it is generally found that a horizontal aerial gives a better value of the ratio signal/noise than a vertical one, usually because the interfering "man-made" static noises from electrical apparatus consist of highly damped high frequency waves which are vertically polarised in many cases.

The horizontal aerial, in common with a vertical aeiral, gives no radiation in the end-on direction in line with the aerial wire [*cf.* paragraph 8]; it will, however, give good reception "end-on" at distances greater than the skip. Moreover, since the horizontal dipole radiates its ground wave horizontally polarised, it follows that a vertical receiving aerial at a short distance from it will not give good reception. For Naval communication work it is usually most desirable that the ground wave should be received at as great a distance as possible, in order that the width of the first "dead space" may be reduced to a minimum; for this reason, vertical dipole aerials are generally preferred. From the point of view of the reception of signals beyond the range of the ground ray, there is little to choose between vertical or horizontal dipoles.

42. **Polar Diagrams for Flat Topped Aerials.**—These are usually of the inverted " L " or " T " type, and are normally used on low frequencies where their length is small in comparison with a quarter wavelength. The radiation from the flat top is horizontally polarised, and that from the vertical part is vertically polarised. The effect of this is that these aerials radiate a vertically polarised wave in the " end-on " direction and a combined horizontally and vertically polarised wave in the " broad-side on " direction. The latter effect simply tilts the angle of the plane of polarisation. For example, if the vertical and horizontal part of the aerial are each one wavelength long, the electric component of the broad-side radiated wave will be oscillating at an angle of approximately 45° to the horizontal. It is interesting to note that if two dipole aerials at the same height are used, one vertical and the other horizontal, and the oscillations in the one lag a quarter cycle behind those in the other, then the broad-side radiated wave will be **circularly polarised.**

The horizontal polar diagram for the " T " shaped aerial is approximately circular so far as the vertically polarised radiation is concerned. That of the " L " shaped aerial is slightly directional in the direction of the horizontal part forming the roof ; it transmits and receives most strongly in the direction tip-to-uplead.

43. **Multiple Aerials, Aerial Arrays.**—Instead of using only one radiating source, it is possible to use various combinations of suitably spaced aerials, all connected to the same source (or receiver). At H/F where both the length of the aerials and their spacings may be made commensurate with the wavelength that is used, it is found that each combination has its own distinctive directional properties ; it should, however, be observed that the same directional properties could be obtained using low frequencies, and only their high engineering cost, due to high masts and large ground areas, have prevented their appearance in engineering practice. At H/F very many combinations have been investigated experimentally, and almost all of them may be said to be an assembly of half wavelength or multiple half wavelength aerials ; in the case of a transmitter, the currents in each unit are phased so that, in conjunction with the spacing between them, a relative increase of field strength is produced in some desired direction. The individual components of the combination, usually called an " aerial array," may, of course, be disposed at any angle although in many cases they are either vertical or horizontal.

For the purpose of a simple classification, an array of aerials along a horizontal line which exhibits marked directional properties in the horizontal plane in a direction at right angles to the line of the array, is usually called a " BROAD-SIDE ARRAY." One that has horizontal directivity along the line of an array is called an end-on or " END-FIRE " array.

For communication work, arrays are designed as much for their vertical directivity as for that in the horizontal plane ; the latter often depends on the total width of the array, usually called the " **aperture,**" while the former is controlled by *stacking* the aerials in tiers, one vertically above the other.

It is proposed to consider briefly the principles and properties of a few of the aerial arrays in practical use.

44. **Principle of Spaced Vertical Aerials.**—In Fig. 45 (*a*), A and B represent two transmitting aerials at a distance apart which can be varied. Suppose that both aerials are transmitting at the same time on the same frequency ; the effect they produce at any distant point C will be the sum of the effects due to the separate aerials.

If the alternating currents in the aerials are in the same phase and the aerials are a half wavelength apart, it is clear that at a point such as D on the line through AB, the wave from A will at every instant be cancelled by the wave from B, because the distance DA differs by a half wavelength from the distance DB. At any points, such as P or P_1 on the line at right angles to AB and through a point O midway between them, the separate waves from A and B will reinforce each other. Thus a maximum of radiation takes place along POP_1 and no radiation takes place along OD, or in the direction opposite to it ; the arrangement constitutes a simple **broad-side array.** At any point C the radiation will be the resultant of a component from A and a component from B, the phases of

the components depending on the difference of time taken by a wave to travel from A to C and from B to C. The horizontal polar diagram of the array can be shown to be a figure-of-eight, as represented in Fig. 45 (*b*).

If the current in aerial B lags 90° on that in A, and if the spacing between them is a quarter wavelength, an entirely different result is produced. The aerial A radiates a wave 90° (or a quarter wavelength) in advance of that radiated from B. Since the separation between A and B is also a quarter wavelength, it follows that by the time the wave from A has reached B, it will be in phase with the wave being radiated therefrom, and will augment the radiation in the direction OD. In the opposite direction OD_1, since the radiation from B starts with a lag of 90° on that from A, the effect of the spacing is that on arrival at A the two radiations are in anti-phase and, accordingly,

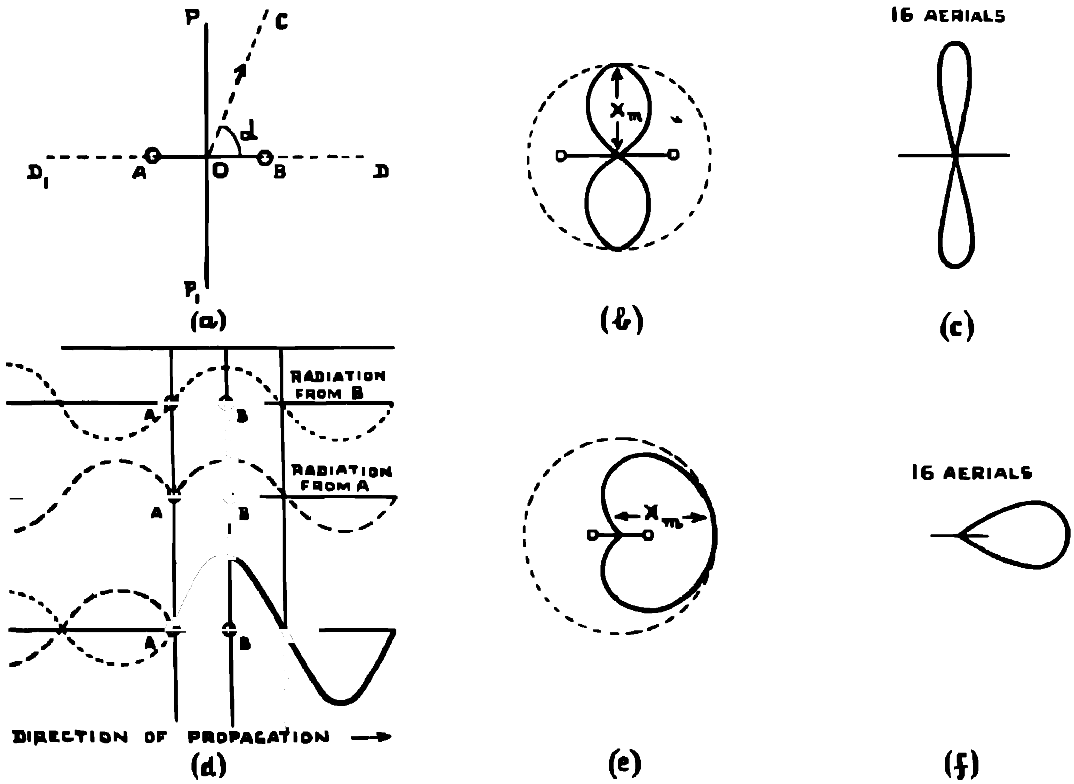


FIG. 45.

cancel each other. This process is represented in Fig. 45 (*d*), and it can be shown that the horizontal polar diagram of the arrangement is a cardioid, as shown in Fig. 45 (*e*). The arrangement is clearly directional along the line of the aerials and is a simple **end-fire array**. The combination also explains the basic theory of the action of **reflector aerials**, a matter which is referred to more fully in paragraph 49.

These are the simplest examples of the use of two transmitting aerials to produce a directional diagram which may be used equally well for transmission or reception. In order to concentrate a larger percentage of the total energy in one direction than is possible with two aerials, it is necessary to employ a number of radiating aerials spaced over a distance of several wavelengths. Fig. 45 (*c*) and (*f*) are drawn for the same two cases using sixteen aerials instead of two.

From these examples it is clear that the currents in the aerials must be maintained in a definite phase relation to each other, though not necessarily all in the same phase. Many different arrangements of aerials have been used to give these directional effects. Theoretically, they all depend upon summing up the separate effects due to each element of current in the array, having due regard to its phase and distance from the point under consideration. It may also be necessary to take into consideration the effect of the conductivity of the sea or earth immediately under the aerials; it often modifies the radiation appreciably.

★45. **Formulae for Horizontal Polar Diagrams of Aerial Arrays.**—Fig. 46 (*a*) represents a plan of " N " aerials spaced " *s* " apart. If the aerials are all radiating in phase and all of the aerial currents are equal, it is clear that there will be reinforcements in the directions PP_1 at right angles to the line of the array, at points where the distance away is large in comparison with the width of the array.

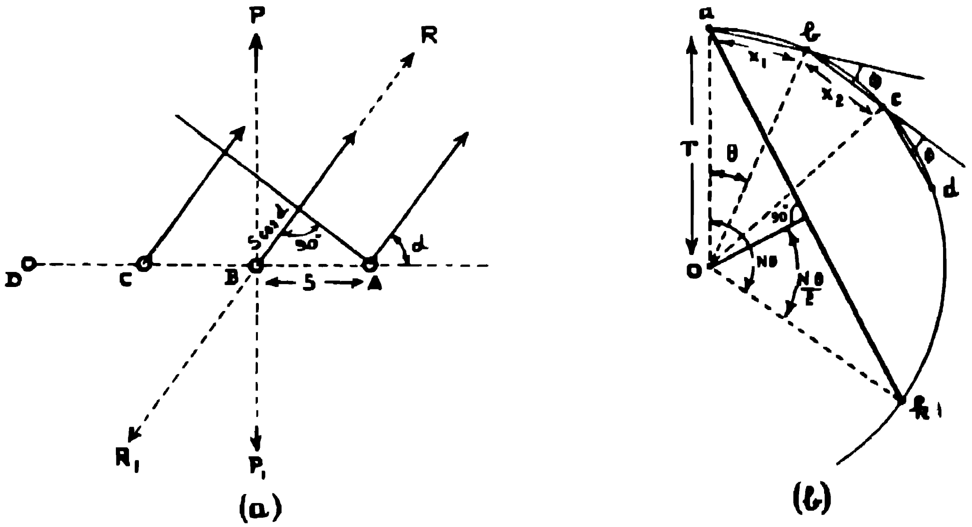


FIG. 46.

If the field produced by one aerial alone is X_1, the maximum field in direction PP_1 will be given by NX_1.

At similar distant points in the RR_1 directions, at angle α to the line of the array, considering two successive aerials A and B, the radiation from B has to travel a further distance $s \cos \alpha$ and accordingly arrives lagging in phase; the geometrical relations are seen in (*a*) above. The value of the phase angle " ϕ " will be a fraction of 2π radians and is given by

$$\phi = \frac{s \cos \alpha}{\lambda} \times 2\pi = \frac{2\pi s \cos \alpha}{\lambda} \quad \ldots\ldots\ldots\ldots \quad (1)$$

This is the constant phase angle between the radiation from any two consecutive aerials in a given direction RR_1. If the aerial currents are all equal, the aerial A produces a field at a distant point sensibly the same in amplitude as that produced by B, but since in general it differs by a phase angle, it follows that the effects can only be added up by the methods of vector addition. For two aerials the graphical representation of this is, in general, a triangle, the closing side representing the resultant; for N aerials, in general, the figure is a polygon, the closing side again being the resultant. Fig. 46 (*b*) shows how to obtain the vector sum given by

$$X = X_1 + X_2 + X_3 + \ldots . X_N$$

an arbitrary direction being chosen for the first vector X_1.

X_1 represents the amplitude of the field due to aerial A

X_2 represents the amplitude of the field due to aerial B,

etc.,

X, the total field in direction RR_1, is the closing side " *ak* " of the polygon.

Now, IF THE PHASE ANGLE BETWEEN THE RADIATION FROM ANY TWO CONSECUTIVE AERIALS IS SMALL, a circular arc may almost be drawn through the ends of the vectors. The vector *ab* is the chord of a circle and subtends an angle θ at the centre point O; it is also the angle between the tangents at *a* and *b*. When the phase angle ϕ is small, the tangent at *b* almost coincides in direction with the chord *ab*, and the angle between the tangents becomes almost equal to the phase angle, that is

$$\theta \doteqdot \phi$$

Angle $aok = \mathrm{N} \doteqdot \mathrm{N}\phi$,

$$\therefore \qquad ak \doteqdot 2r \sin\frac{\mathrm{N}\theta}{2} \doteqdot 2r \sin\frac{\mathrm{N}\phi}{2} = \mathrm{X},$$

where r = radius of the circle,

$$\text{and} \qquad ab = 2r \sin\frac{\theta}{2} \doteqdot 2r \sin\frac{\phi}{2} = \mathrm{X}_1,$$

where *ak* represents the total field "X" in direction R, and

ab represents the field X_1 produced by one aerial only.

The maximum value of the field is in direction P and is given by $\mathrm{X}_m = \mathrm{N}\mathrm{X}_1$.

Now X_m must be greater than X, since arc "*ak*" must exceed the chord "*ak*."

$$\text{We have} \qquad \frac{\mathrm{X}}{\mathrm{X}_m} = \frac{2r\sin\frac{\mathrm{N}\phi}{2}}{2\mathrm{N}r\sin\frac{\phi}{2}} = \frac{\sin\frac{\mathrm{N}\phi}{2}}{\mathrm{N}\sin\frac{\phi}{2}}$$

$$\text{Usually written} \qquad \mathrm{X} = \mathrm{X}_m \frac{\sin\frac{\mathrm{N}\phi}{2}}{\mathrm{N}\sin\frac{\phi}{2}} \qquad \ldots\ldots\ldots\ldots \quad (2)$$

It is usual to take the value of X_m to be unity, since in plotting polar curves it is, most often, the values relative to the maximum value which are of interest. The polar radius is then given by—

$$\mathrm{X} = \frac{\sin\frac{\mathrm{N}\phi}{2}}{\mathrm{N}\sin\frac{\phi}{2}} \qquad \ldots\ldots\ldots\ldots \quad (3)$$

and from (1)

$$\mathrm{X} = \frac{\sin\left(\frac{\mathrm{N}\pi s\cos\alpha}{\lambda}\right)}{\mathrm{N}\sin\left(\frac{\pi s\cos\alpha}{\lambda}\right)} \qquad \ldots\ldots\ldots\ldots \quad (4)$$

In strict analysis, when ϕ is large, the vectors cannot be said to lie along the arc of a circle, and the reasoning must be different. In spite of various approximations which have been made, equation (3) represents a result which is valid for all values of ϕ, and it may be adapted to include cases where the aerial currents are out of phase. If the total width of the aerial array is a whole number of wavelengths, the field at a distant point due to half of the aerials will exactly cancel that due to the other half, and the radiation along the line of the aerials will be zero.

The positions of the minima, or extinction directions, may be calculated by equating the numerator of (2) to zero, π, 2π, etc. The positions of the maxima can be found by applying the methods of the differential calculus to equation (2).

The use of the general formula No. 4 will now be made clear by reference to certain special cases.

Case 1.

$$\mathrm{N} = 1.$$

From (3) we have

$$\mathrm{X} = \frac{\sin\frac{\phi}{2}}{\sin\frac{\phi}{2}} = 1.$$

The polar diagram is a circle and $\mathrm{X}_m = \mathrm{X}_1$.

Case 2.

$$\mathrm{N} = 2,\ s = \frac{\lambda}{2},\ \text{currents in phase.}$$

From (3) $$X = \frac{\sin\phi}{2\sin\frac{\phi}{2}} = \cos\frac{\phi}{2} = \cos\left(\frac{\pi\cos\alpha}{2}\right)$$

since from (1), $\phi = \pi\cos\alpha$.

MINIMUM values $$\frac{\pi\cos\alpha}{2} = \pm\frac{\pi}{2}$$

i.e., $$\cos\alpha = \pm 1,\quad \alpha = 0^\circ,\ 180^\circ.$$

MAXIMUM values $$\frac{\pi\cos\alpha}{2} = 0,$$

i.e., $$\cos\alpha = 0,\ \alpha = 90^\circ,\ 270^\circ.$$

This produces the figure-of-eight diagram of Fig. 45 (d) where $X_m = 2X_1$.

Case 3.

$$\mathrm{N} = 2,\ s = \frac{\lambda}{4},\ \text{currents } \frac{\pi}{2} \text{ out of phase.}$$

From (1) $$\phi = \frac{\pi\cos\alpha}{2} - \frac{\pi}{2},$$

the current in aerial B leading that in A.

From (3) $$X = \cos\frac{\phi}{2} = \cos\left(\frac{\pi\cos\alpha}{4} - \frac{\pi}{4}\right).$$

MINIMUM values $$\frac{\pi\cos\alpha}{4} - \frac{\pi}{4} = \pm\frac{\pi}{2}$$

i.e., $$\cos\alpha = -1,\ \alpha = 180^\circ.$$

MAXIMUM values $$\frac{\pi\cos\alpha}{4} - \frac{\pi}{4} = 0,$$

i.e., $$\cos\alpha = 1,\ \alpha = 0^\circ.$$

This produces the cardioid diagram of Fig. 45 (*e*), where $X_m = 2X_1$.

Case 4.

$$\mathrm{N} = 4,\ s = \frac{\lambda}{2},\ \text{currents in phase.}$$

From (3) $$X = \frac{\sin 2\phi}{4\sin\frac{\phi}{2}} = \cos\phi\cos\frac{\phi}{2}$$

$$= \cos(\pi\cos\alpha)\cos\left(\frac{\pi\cos\alpha}{2}\right), \qquad \text{(5)}$$

since from (1), $\phi = \pi\cos\alpha$.

MINIMUM values $$\pi\cos\alpha = \pm\frac{\pi}{2}$$

i.e., $$\cos\alpha = \pm\tfrac{1}{2},\ \alpha = 60^\circ,\ 120^\circ,\ 240^\circ,\ 300^\circ.$$

also for $$\frac{\pi\cos\alpha}{2} = \pm\frac{\pi}{2}$$

i.e., $$\cos\alpha = \pm 1,\ \alpha = 0^\circ,\ 180^\circ.$$

Differentiating (5) and equating to zero in the usual way shows that the first subsidiary maximum occurs when $\alpha \doteqdot 42°$, and that the principal maximum is at right angles to the lines of the array. Fig. 47 (a) shows the form of the polar diagram, where $X_m = 4X_1$ for the case where $\alpha = 090°, 270°$.

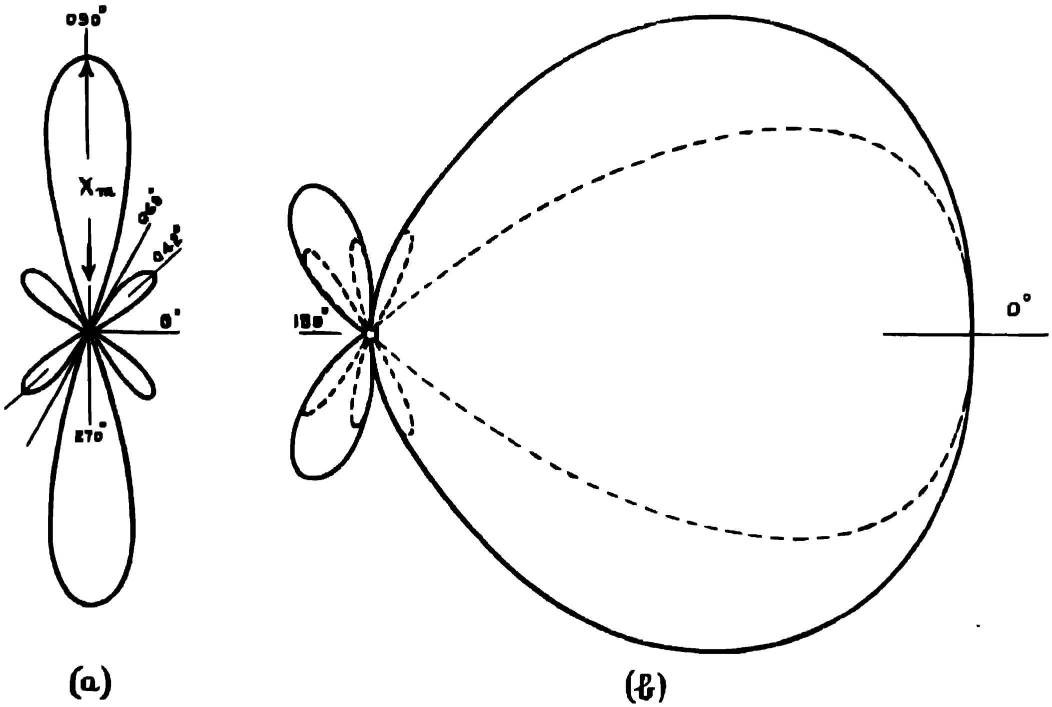


FIG. 47.

Case 5.

$$N = 4,\ s = \frac{\lambda}{4},\ \text{currents successively } \frac{\pi}{2} \text{ out of phase.}$$

From (3)
$$X = \cos\phi \cos\frac{\phi}{2} \quad \ldots\ldots \text{and as in Case 2.}$$

$$= \cos\left(\frac{\pi\cos\alpha}{4} - \frac{\pi}{4}\right)\cos\left(\frac{\pi\cos\alpha}{2} - \frac{\pi}{2}\right). \quad \ldots\ldots\ldots (6)$$

MINIMUM values
$$\frac{\pi\cos\alpha}{4} - \frac{\pi}{4} = \pm\frac{\pi}{2},$$

i.e.,
$$\cos\alpha = -1,\ \alpha = 180°.$$

also for
$$\frac{\pi\cos\alpha}{2} - \frac{\pi}{2} = \pm\frac{\pi}{2}$$

i.e.,
$$\cos\alpha = 0,\ \alpha = 90°,\ 270°$$

MAXIMUM values
$$\frac{\pi\cos\alpha}{2} - \frac{\pi}{2} = 0.$$

i.e.
$$\cos\alpha = 1,\ \alpha = 0°.$$

Differentiating (6), followed by the usual treatment, shows that equal subsidiary maxima occur for $\alpha \doteqdot 115°$, $245°$; Fig. 47 (*b*) shows the form of the polar diagram, the dotted curve showing the result of using eight aerials instead of four, the full treatment of which will not be given here. In that case, however, the formula for field strength reduces to—

$$X = \cos\frac{\phi}{2} \cdot \cos\phi \cdot \cos 2\phi$$

where

$$\phi = \frac{\pi\cos\alpha}{2} - \frac{\pi}{2}$$

46. **The Bruce Type of Mesny Array.**—The Bruce aerial represents one of the many methods of setting up a line of vertical aerials and energising them in the same phase in order to produce a broadside array. It is also known as a " **zig-zag** " aerial by reason of the nature of its appearance which is represented in Fig. 48 The array is supported with its plane vertical by slinging it between triatics. The sides of the zig-zag are everywhere a quarter wave in length, with the exception of the two ends which are an eighth of a wavelength long. The overall length of the array is usually between 5 to 10 wavelengths, in order to give the requisite horizontal directivity.

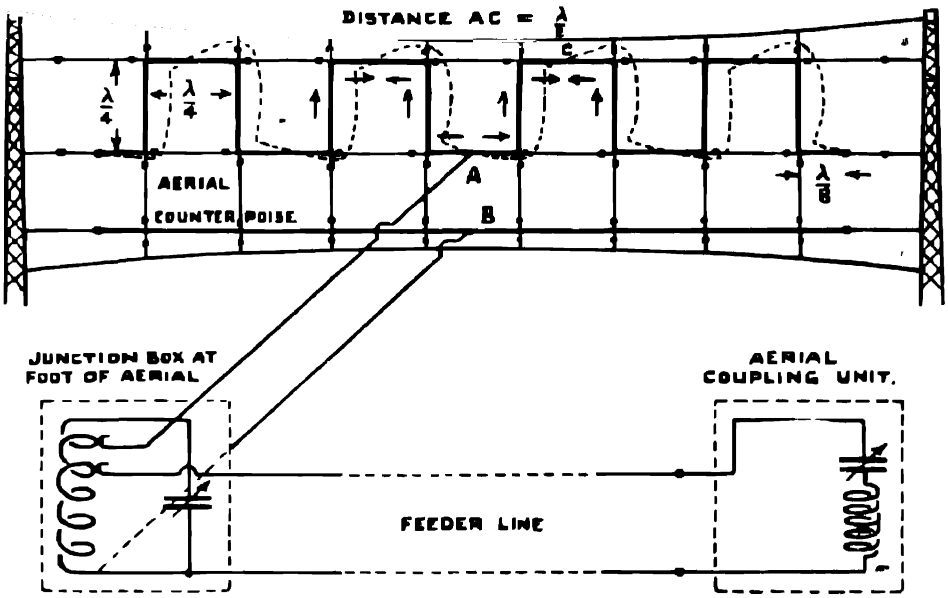


FIG. 48.

The array is used chiefly for reception but its action is more easy to understand if we assume that it is being used for transmission. The system may be regarded as a centre fed aerial, the length being long in comparison with the wavelength in use. Stationary waves will be set up, with a current node at each end and at the centre of each horizontal member, and a current maximum at the centre of each vertical member. The effect is that the radiation from the two halves of the horizontal members mutually cancel each other, and that the whole constitutes a number of vertical aerials radiating in the same phase. It will be noted that this is entirely due to bending the wire forming the aerial into a zig-zag shape; the effect is somewhat similar to that of the phasing coils in a Franklin " **uniform** " aerial.

The system is " voltage fed," since it is energised at the voltage anti-node which occurs at the centre, B. The feed is usually effected by means of a coupling unit between B and a counterpoise, or the earth. Instead of the straight wire counterpoise shown in the diagram, one may use a lower

zig-zag aerial arranged to be the mirror image of the upper one. This increases the " pick-up " when the array is used for reception, and also gives that increased directivity in the vertical plane which is always the result of stacking.

This array gives excellent horizontal directivity, provided that the **aperture** is five wavelengths or more. Moreover, the directional properties are not a direct function of λ, and are not impaired if the aerial is used for some frequency slightly different from that for which it is designed. This is to be contrasted with end-on arrays, the directional properties of which are a function of the wavelength, and hence of the spacing between the various out-of-phase aerials.

In the Service, this simple array is used for reception at some shore stations engaged in working certain point-to-point services. The plane of the array is arranged to be at right angles to the direction of the signals that it is desired to receive. If the incoming electric field is vertically polarised, a standing wave system will be set up similar to that which is produced when the aerial is used as a transmitter. The array is somewhat selective, and for Naval work it is convenient to keep spare arrays designed to work on certain other emergency frequencies.

Since the array is fed at a high impedance point, it behaves as a rejector circuit. It is necessary to set up a carefully matched impedance system between the receiver coupling unit, the feeder lines, and the aerial. This impedance matching is achieved in the ordinary way by using an auto-transformer coupling at each end of the feeder line. The process of tuning to ensure that the feeder lines are non-reflective is, naturally, an elaborate technical matter. It is usually done by energising the aerial as a low power transmitter by means of a local oscillator. When the various units are in tune, the feeder tapping point to the inductance in the junction box at the foot of the aerial is adjusted until three or more milliammeters, spaced $\lambda/6$ apart in the vicinity of the box, all show the same reading. When this is the case, the feeders will have been brought to the necessary non-resonant condition (*Cf.* paragraph 38, VI) ; the high impedance aerial will have been matched to a relatively low impedance feeder line, the impedance of the latter being matched to the receiver to which it is coupled.

This array has the ordinary two-sided or bi-directional figure-of-eight characteristic. It may be given the uni-directional properties of a " **beam system,**" by mounting a similar zig-zag aerial at a distance of a quarter wavelength behind it. This second aerial need not be energised, and will operate as a reflector in a manner which will be described more fully in paragraph 49.

47. **Marconi-Franklin Series Phase Aerial.**—This is a product of the Marconi Company, and was first described fully in 1933 It is here cited as an example of an **end-on-array,** the line of the aerial pointing in the direction of the station from which it is desired to receive or to which it is hoped to transmit.

The basic principles have already been described in paragraph 44. The general features include a uni-directional characteristic which provides " beam " transmission or reception (paragraph 49) without the need of a reflector aerial. It requires low aerial masts, made preferably of wood or any other non-absorber of energy, and it is now regarded as an alternative to the older Marconi beam system, using the high Franklin uniform aerials.

The system is shown diagrammatically in Fig. 49 (*a*), and can best be understood by considering its action as a transmitter, although the process is entirely a reversible one. Travelling waves are fed upwards through a non-radiating feeder to the end A of the aerial. From there they travel along the aerial to Q, down a portion of non-radiating feeder to a resistance, equal to the surge impedance of the system, to absorb any residual energy. The whole action of the system depends upon *complete* absorption of travelling energy at the end Q ; at this point there is no objection to using a mast made of steel. The dotted curves in Fig. 49 (*a*) represent a travelling current wave at an instant of time ; assuming no attenuating losses, it is similarly represented in Fig. 49 (*b*).

The two limbs of any loop are sufficiently close together in space that—as regards radiation—they may be considered as coincident and hence replaceable by a single vertical wire on which there are two waves of equal amplitude travelling in opposite directions. Stationary waves will, therefore, be set up (R. 14), the nodes being situated at the points B, E, H, M, etc., since at those points there

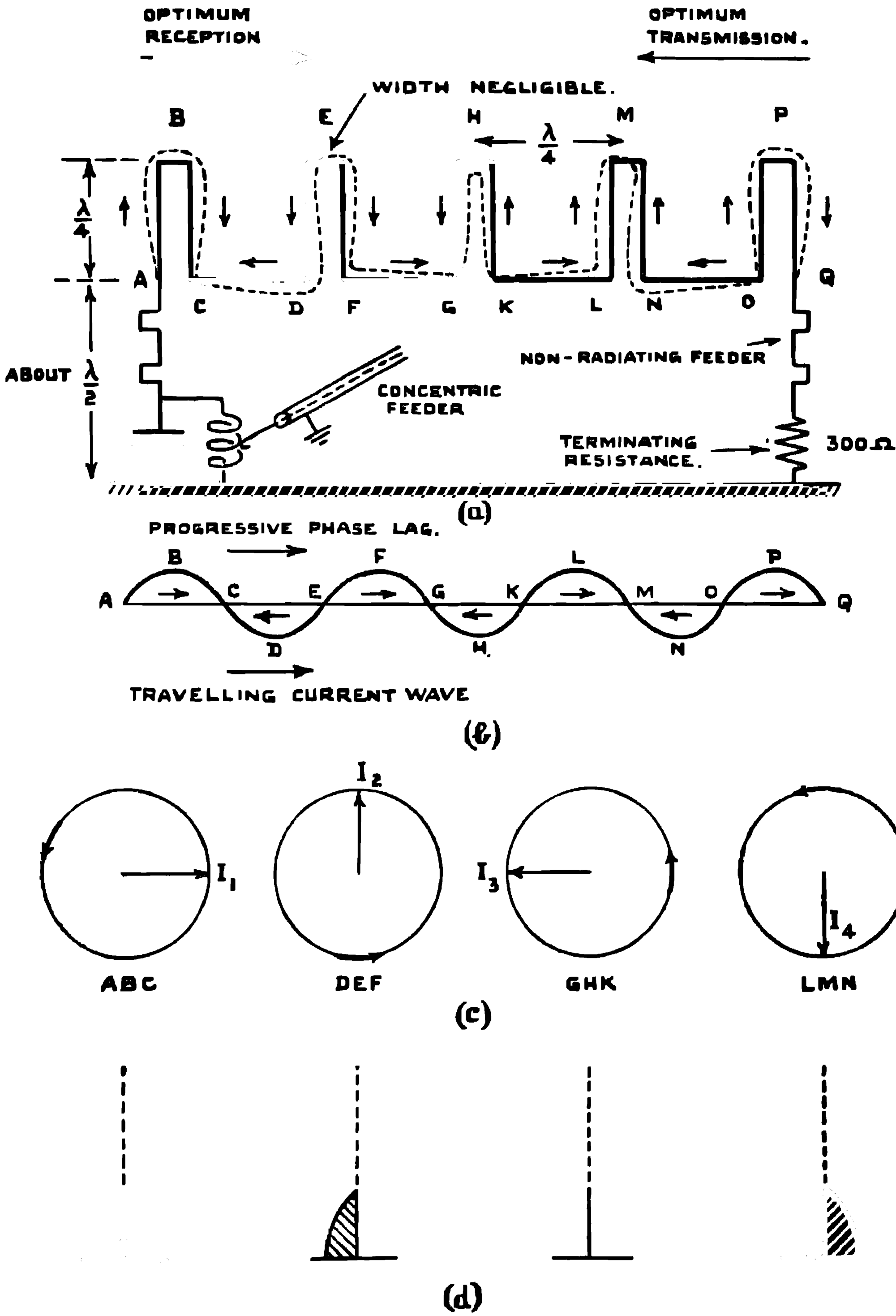

OPTIMUM
RECEPTION
OPTIMUM
TRANSMISSION.
WIDTH NEGLIGIBLE.
B
E
H
M
P
$\frac{\lambda}{4}$
$\frac{\lambda}{4}$
A
C
D
F
G
K
L
N
O
Q
ABOUT $\frac{\lambda}{2}$
NON-RADIATING FEEDER
CONCENTRIC
FEEDER
TERMINATING
RESISTANCE.
300Ω
(a)
PROGRESSIVE PHASE LAG.
B
F
L
P
A
C
E
G
K
M
O
Q
D
H.
N
TRAVELLING CURRENT WAVE
(b)
I_2
I_1
I_3
I_4
ABC
DEF
GHK
LMN
(c)
(d)

FIG. 49.

will always be two equal currents flowing in opposite directions. Each loop will, therefore, radiate in the same way as a single quarter wavelength aerial carrying a standing wave.

Consideration of the direction of the arrows in Fig. 49 (*a*) and (*b*) shows, moreover, that members of this line of aerials are not radiating in the same phase at the same instant of time. Fig. 49 (*c*) and (*d*) show diagrammatically the relative phase of each of the aerials, the vectors of Fig. 49 (*c*) indicating the progressive phase difference of 90° between successive radiators. At the instant of maximum radiation—that chosen for the diagram—it wil ,e seen that only each alternate aerial is radiating, DEF, LMN, etc.; the current in DEF *leads* tnat in ABC by 90°, and so on down the line from the end A. [*Note.*—A lag of 270° is electrically equivalent to a lead of 90°.]

Although each loop is equivalent to a quarter wavelength aerial, there is an important difference; it can be shown that the effective radiation current is doubled, increasing the power radiated and the radiation resistance by four times. From this it follows that the loops ABC, DEF, etc., may now be considered to be replaced by single wires spaced a quarter of a wavelength apart, having equal R.M.S. current in each, but with a phase difference of 90° between adjacent wires, the phase lagging progressively from A to O. These are the required conditions for an end-on array, the direction from O to A being that in which reinforcement occurs. Considering vector I_2, it represents an aerial radiating a wave 90° ahead of I_1; since it is spaced by a quarter wavelength from the aerial represented by I_1, its effect at that point will be equivalent to a wave arriving in phase. This reinforcement in the forward direction between the aerials corresponding to vectors I_1 and I_2, is represented by rotating the latter backwards through 90°. In the backward direction from A to O, since the wave from the equivalent aerial at ABC starts with a lag of 90°, by the time it arrives at the aerial equivalent to DEF it will be lagging by another 90°, and will therefore arrive exactly in anti-phase. Successive pairs of aerials provide mutual cancellation in the backward direction.

The diagram shows the aerial being fed and terminated through non-radiating feeders; these may be designed in various ways, and this is an essential feature in order to avoid an all-round pick-up at either of these points which would spoil the shape of the polar diagrams. The terminating resistance is usually of the order of 300 ohms; when the length amounts to four wavelengths or more, it is usually found that no termination is necessary, the travelling energy being wholly dissipated in the wire. At the supply end, the aerial is joined to a feeder system through an impedance matching circuit of the usual form.

Increased directivity in the horizontal plane may be obtained by adding further parallel elements in broadside array; it is quite common to use four of these elements, fed in pairs from a branched feeder line similar to that described for the Marconi beam system. The theoretical polar diagram for a single element of four loops or equivalent aerials is shown in Fig. 47 (*b*). In the vertical plane, the system gives chiefly low angle radiation at an angle of about 10°. The aerial gain is very good, and with an array of four elements it is possible to achieve a discrimination between front and back reception of about 25*db.*, equivalent to a power ratio of 316 : 1.

For transmitting purposes, it has been found that satisfactory results are produced with aerials four wavelengths long, having 17 complete loops. For receiving purposes, the aerials may be shorter, two wavelengths, or nine complete loops being a satisfactory figure. It may be noted that for complete cancellation of radiation in a backward direction, from theoretical considerations it would appear that there should be an even number of radiating loops. Since the horizontal polar diagram of an element with (say) 17 loops, includes a large number of small lobes in the two back quadrants, it is not a matter of much importance if there is also a small "tail" exactly at the back. In the case of reception, interference is liable to come from any random direction, and in practice both odd and even numbers of radiating loops are used.

In connection with the horizontal members CD, FG, etc., of the system, it will be noted that the currents are successively 270° apart, and that the current in FG is 90° ahead of that in CD. At a distant point along the direction normal to the line of the array, the radiation from successive horizontal portions will suffer partial mutual cancellation, and the nett effect will be small if the system contains an odd number of radiating loops. At distant points along the line of aerials in the forward direction, since CD and FG are spaced by about a quarter wavelength, some reinforcement will occur for the reason which has been explained in the case of the vertical members. It is clear that this radiation is horizontally polarised.

★**Mathematical Note.**—The current distribution at any point in the system, at any instant of time, may be obtained analytically as follows :—

Measuring distances from A, the points A, D, G, L, O are each distant $3n\lambda/4$ from A, where $n = 0, 1, 2, 3,$ etc.

Similarly, points C, F, K, N, etc., are each distant $(3n + 2)\lambda/4$ from A, where $n = 0, 1, 2, 3,$ etc.

Any two corresponding points X and Y, each at the same height d from the base of the two sides forming a loop such as LM and NM respectively, are hence represented by distances—

$$(3n\lambda/4 + d) \quad \text{and} \quad [(3n + 2)\lambda/4 - d] \quad \text{from A.}$$

Assume that the instantaneous current at any point distant x from A due to a travelling wave moving in the direction ABCD . . . Q is given by the expression

$$= \mathcal{I} \sin\left(\omega t - \frac{2\pi x}{\lambda}\right) \quad \ldots\ \ldots \qquad [cf.\ \text{T4}].$$

This is more usually written—

$$i = \mathcal{I} \sin 2\pi\left(\frac{t}{T} - \frac{x}{\lambda}\right) \quad \ldots\ldots \quad \text{since } \omega = \frac{2\pi}{T}.$$

$$\text{Hence current at} \qquad X = \mathcal{I} \sin 2\pi\left(\frac{t}{T} - \frac{3n\lambda/4 + d}{\lambda}\right)$$

$$\text{and current at} \qquad Y = \mathcal{I} \sin 2\pi\left(\frac{t}{T} - \frac{(3n + 2)\lambda/4 - d}{\lambda}\right).$$

From the point of view of *radiation*, LM and NM—and other similar loops—may be considered coincident in space ; but as regards *current*, the side NM is reversed relative to LM, and hence the effective radiation current at the coincident point X/Y in space is given by the difference—

$$\mathcal{I} \sin 2\pi\left(\frac{t}{T} - \frac{3n\lambda/4 + d}{\lambda}\right) - \mathcal{I} \sin 2\pi\left(\frac{t}{T} - \frac{(3n + 2)\lambda/4 - d}{\lambda}\right)$$

$$= \mathcal{I} \sin\left[2\pi\left(\frac{t}{T} - \frac{3n}{4}\right) - \frac{2\pi d}{\lambda}\right] - \mathcal{I} \sin\left[2\pi\left(\frac{t}{T} - \frac{3n}{4}\right) - \frac{2\pi}{\lambda}\left(\frac{\lambda}{2} - d\right)\right]$$

$$= 2\mathcal{I} \cos\left[2\pi\left(\frac{t}{T} - \frac{3n}{4}\right) - \frac{\pi}{2}\right] \sin\left[\frac{\pi}{2} - \frac{2\pi d}{\lambda}\right]$$

$$= 2\mathcal{I} \cos\left[2\pi\frac{t}{T} - \frac{\pi}{2}(1 + 3n)\right] \sin\left[\frac{\pi}{2} - \frac{2\pi d}{\lambda}\right]$$

$$= 2\mathcal{I} \cos\frac{2\pi d}{\lambda} \cos\left[2\pi\frac{t}{T} - \frac{\pi}{2}(1 + 3n)\right] \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

The value of d is, of course, restricted to the range from zero to $\lambda/4$. Each coincident pair of aerials therefore behaves as if it had the standing wave distribution of current of a quarter wave aerial, given by $2\mathcal{I} \cos 2\pi d/\lambda$. Moreover, from (1) we see that the effective current at any instant in any pair, say LM/NM, **lags** behind that in the preceding pair GH/KH by $\frac{3\pi}{2}$, *i.e.*, it **leads** by $\frac{\pi}{2}$. Radiation will hence be reinforced in the direction QA and cancelled in direction AQ. The relative phase of the currents in the aerials at an instant of time may also be seen by putting $t = 0$ and taking successive values of $n = 0, 1, 2,$ etc., in equation (1).

The power radiated by one loop will vary as $(2\mathcal{I})^2$; the radiation resistance will therefore be four times the value of that of a simple $\lambda/4$ aerial.

48. Advantages of Arrays, Array Gain.—For transmitting purposes the use of an array prevents some of the wasteful radiation of energy in unwanted directions that characterises an "all-round" or broadcast aerial ; the energy radiated is more usefully applied. This means that

an array can produce a given field strength at a distant point using much less power than an equivalent single aerial; or, alternatively, if the same power is applied to both an array and a single aerial, the former will produce a greater field strength along the optimum direction. This fundamental aspect of the advantage of an array can be given precise numerical form.

★(*a*) AN ARRAY, AND A SINGLE EQUIVALENT AERIAL.—Assume N similar spaced aerials, each with current I and of radiation resistance R, the phasing and spacing being such that in some directions reinforcement occurs, giving a field strength N times that from one aerial alone. Neglecting the effect of mutual coupling between aerials—

Total power radiated by array $P_N = NRI^2$. .... (1)

Optimum field strength produced $X_N = K\,(NI)$,

where K is a constant (paragraph 28).

Power radiated by one array element $P_1 = RI^2$

Field strength produced by an element $X_1 = KI$.

In order to produce a field strength of value X_N using one element only, it is clearly necessary to use an aerial current "I_o" which is N times that in one element of the array, but in that case, the power radiated by the Single Equivalent Aerial (S.E.A.) is given by—

$$P_o = RI_o^2 = R\,(NI)^2, \qquad (2)$$

and this is greater than P_N, hence—

$$\text{Array improvement factor} = \frac{P_o}{P_N} = \frac{R\,(NI)^2}{NRI^2} = N.$$

This means that an array of 8 aerials can produce a given field strength using one-eighth of the power required by one S.E.A., and with one-eighth of the aerial current.

(*b*) AN ARRAY AND A SINGLE STANDARD COMPARISON AERIAL, EACH SUPPLIED WITH THE SAME POWER. We may consider that this refers to the use of a given source of power either with a single aerial, or with an array of (say) the zig-zag type.

With assumptions as before—

$$P_N = NRI^2 = RI_o^2,$$

since $P_N = P_o$, and I_o is the requisite current in the single aerial.

$$\therefore \quad I_o = I\sqrt{N}.$$

Then $$X_N = K\,(NI), \text{ and } X_1 = KI_o = K\,(I\sqrt{N}),$$

$$\therefore \quad \frac{X_N}{X_1} = \sqrt{N}.$$

This means that if a given source of power is divided between (say) two aerials in array, the field strength in the optimum direction will be $\sqrt{2}$ times that produced by a single standard comparison aerial.

Case (*b*) provides a simple experimental method of comparing a directive array with a single aerial, though it is probable that case (*a*) suggests a more comprehensive basis. In the latter case the ratio $\frac{P_o}{P_N}$ gives the power which must be supplied to a single equivalent aerial divided by the power that must be supplied to the aerial array, in order to produce a given field strength in some desired direction. If the array improvement factor is expressed in decibels it is usually called the "gain" of the array.

It is possible to arrive at an estimate of the value of the ratio $\frac{P_o}{P_N}$ if the form of the solid polar diagram to the array and of the single equivalent aerial is known. From paragraph 8 it should be clear that the power radiated in each case is proportional to the total volume of the solid polar diagram. In some cases this may be taken to be proportional to the area of the horizontal polar diagram, which is the cross section of a solid polar diagram in a horizontal plane. In Fig. 45 (*a*) and (*e*), the dotted curve shows the circular graph of radius X_N, usually called the unit circle, and represents the horizontal polar diagram of the single equivalent aerial. The polar graph of

the array occupies a smaller area, a fact which shows that, as a transmitter, such a directive array uses less power. From theoretical considerations polar diagrams can be plotted for a number of different cases ; we may say that, provided the radius factor reaches its maximum value, the array having the polar graph of smallest relative area has the greatest horizontal directive gain.

$$\text{Array improvement factor} = \frac{\text{Area of circle}}{\text{Area of relative horizontal polar diagram}}$$

The horizontal polar diagram of the array, whether theoretically or practically obtained, must take account of mutual effects between component aerials. Accordingly, this method of estimating gain eliminates one of the assumptions first stated and gives a better estimate of the advantage of the array.

It was on this basis of comparison that Foster showed that, for two aerials radiating in phase, the best spacing was $0{\cdot}609\lambda$, in order to produce a diagram having the smallest relative area, namely $1 : 0{\cdot}298$. With that spacing the diagram is similar to Fig. 45 (*b*), but with two small secondary lobes along the 0° and 180° directions. When the spacing is $0{\cdot}5\lambda$, the ratio of the areas is $1 : 0{\cdot}3479$ Diagrams for cases where the phase difference is a quarter of a period, all have an area ratio of $1 : 0{\cdot}5$. On the same basis, with 16 aerials in phase, the best spacing is $0{\cdot}88\lambda$, the relative areas then being $1 : 0{\cdot}025$.

On this basis of estimation, for the case of two aerials radiating in phase and spaced half a wavelength apart, we have

$$\frac{P_o}{P_N} = \frac{1}{0{\cdot}3479} = 2{\cdot}88,$$

or using decibels

$$\text{Array gain} = 10 \log_{10}\left(\frac{P_o}{P_N}\right) = 4{\cdot}6 \text{ dbs.}$$

From the point of view of reception, the use of an aerial array means the relative reduction of received energy from random directions, especially when the array is "one-sided." Static noise is also reduced and a high signal/noise ratio is maintained by decreasing the denominator of that ratio.

49. **Beam Transmission.**—Broadside arrays having figure-of-eight directional characteristics, may be made uni-directional in behaviour if a suitable reflector system is available which will concentrate the radiated energy in a 'forward' direction. It is clear that the use of a uni-directional beam, with proper concentration of energy in both zenithal and horizontal planes, will produce still further power economy in the case of transmitting arrays, and provide for still better reception in the case of a receiving array.

The basic principle of a reflector aerial was referred to in paragraph 44, and is illustrated in Fig. 45 (*d*) and (*e*). It is the same as that which accounts for the action of an end-fire array which has, essentially, two aerials spaced by a quarter wave-length, with currents 90° out of phase. The uni-directional property of an end-on array can be combined with the highly directive characteristic of a broadside one, by arranging a row of reflector aerials at a distance of a quarter wavelength behind a similar row of aerials radiating a wave 90° out of phase with that coming from the reflector row. Reinforcement and cancellation will occur in mutually opposite directions at right angles to the line of the array

The two rows of aerials may be similar in all respects. It is not necessary that both rows should be separately excited ; usually one is directly excited and the other is indirectly impulsed by energy received from the directly energised line of aerials. The indirectly impulsed aerials are usually referred to as "**reflector aerials.**" By the time a wave radiated from an aerial in the directly excited row has travelled to the reflector, the aerial voltage has changed by 90°. By Lenz's law, the back E.M.F. in the reflector lags 180° behind the cause which produces it, implying that the reflector back E.M.F. lags 270° behind the aerial voltage. Electrically, the lag of 270° is equivalent to a lead of 90°, and the reflector may be regarded as equivalent to a directly energised aerial, radiating a wave 90° (or $\lambda/4$) in advance of the main aerial wave. By the time the wave has reached the main aerial row, it will be in phase with the wave being radiated therefrom ; it will therefore augment the forward radiation, and oppose the backward radiation towards the reflector row.

This principle may be applied to many different aerial arrays, and reference has already been made to its use in connection with the Bruce array. Its use in connection with the Marconi beam system is described in the next paragraph.

When the two rows of aerials are precisely similar, reversal of the beam is made possible by interchanging their functions.

At frequencies of the order of 30,000 kc./s. a beam can be produced in the same way as in a searchlight, by using the focussing properties of a parabolic reflector. At very high frequencies the latter can be made of sheet metal, but at the frequency in question it usually consists of a number of vertical wires separated about an eighth of a wavelength apart and arranged in a parabolic arc; the transmitter is usually a single wire vertical aerial placed at the focus of the parabola and a distance of either $\lambda/4$ or $3\lambda/4$ from the reflector wires. A rotating beam could be made by revolving such a reflector system about the transmitting aerial as axis. One of the first of such **rotating beacon systems** was installed at Inchkeith by the Marconi Company, and although the width or aperture of the parabola was only about 2 wavelengths, the beam was sufficiently sharp to give an angular accuracy of about 3° at a range of 10 miles.

If the currents in the line of directly excited aerials are not all in phase, the beam will not be at right angles to the line of the array; if the current supplied to one aerial is slightly out of phase with that supplied to its neighbour before it in the line, it may be shown that the effect is to swing the beam round through some angle which will depend upon the amount of the current mis-phase between successive aerials. In practice, it is never more than about 10°. Sometimes it is necessary deliberately to take advantage of this fact, mis-phasing the currents supplied so as to get the maximum radiation in some other direction. It is clear that this can easily be done where the various aerials are separately fed, as in the case of the Marconi beam system which is described later; it would not be easy to arrange in a "series fed" array of the zig-zag type.

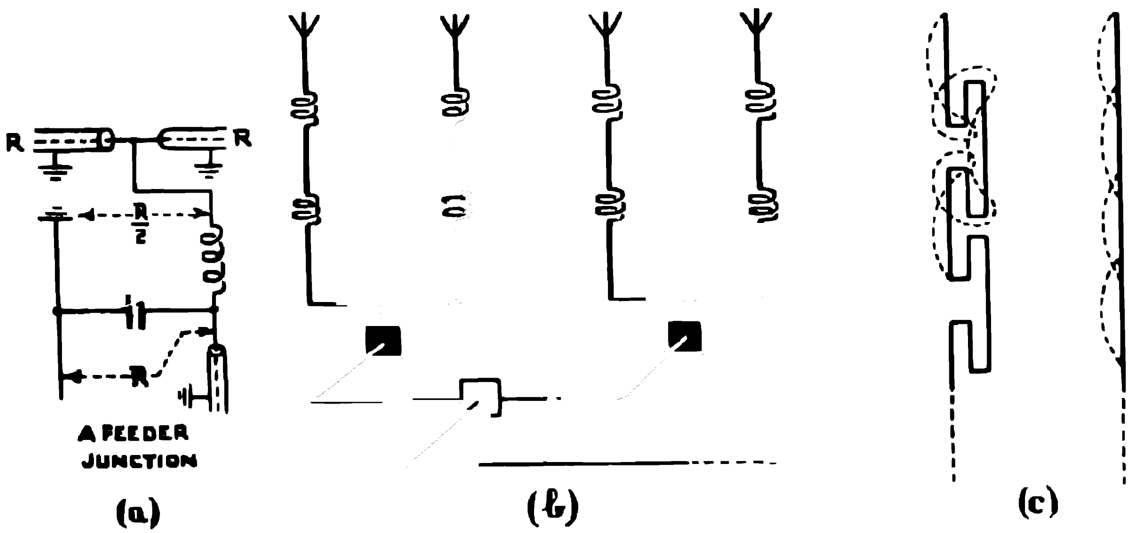


FIG. 50.

50. **Marconi Beam System.**—The first Imperial wireless beam service, operated on the Marconi system, was opened in October, 1926, between Great Britain and Canada. It marked the culminating point of 10 years of H/F experimental work, largely due to Senator Marconi, and was a turning point in wireless history in that it marked the end of the development of high power L/F transmitting stations, of which Rugby may be said to be an example.

The system consists of a uni-directional broadside array composed of a line of directly excited

Franklin " uniform " aerials [paragraph 40, Fig. 43 (*k*) and (*l*)], spaced usually a half wavelength apart, with another parallel line of reflector aerials situated a quarter wavelength behind, having aerials spaced a quarter wavelength apart. The directly excited "in-phase" aerials are energised in pairs from a branched concentric feeder system, shown diagrammatically in Fig. 50 (*a*) and (*b*).

The rows of aerials are several wavelengths in height, and alternate half wavelength suppression is carried out in an improved manner, the older phasing coils having been replaced by a "folded back" aerial. This is shown diagrammatically in Fig. 50 (*c*), and represents an approach to the ideal ***uniform current*** aerial.

Reflector aerials are usually built up of insulated half wavelength sections.

The system produces a beam inclined upwards at an angle of about 15° to the horizontal, having a divergence of about 11° depending upon the aperture of the array.

The use of a multiple feeder system makes it easy to mis-phase the currents supplied to the aerials in line. It is sometimes done by altering the equality of the length of the subsidiary feeder lines. In the Marconi system, correct impedance matching and control of the current phase is simultaneously done by arranging a simple parallel circuit—Fig. 50 (*a*)—at each bifurcation of the feeder. Reference has already been made to the use of this mis-phasing in order to swing the beam through some small angle in case of need.

With a given aerial system, it is possible to cover a frequency band some 15 per cent. in width without serious loss of efficiency. This is due to the high radiation resistance of the array, which, in turn, is partly due to the use of twice as many reflector aerials as energised ones.

51. **Horizontal Arrays, Empire Broadcasting.**—It has been observed in paragraph 41 that a horizontal half-wave aerial, situated a quarter wavelength above the earth, is rich in high angle radiation This may largely be countered by arranging an array of half wavelength sections, in which one is stacked above the other, separated by a half wavelength and excited in phase. At a point vertically above the array the radiation will cancel out ; at a distant point in the horizontal plane at right angles to the line of the array, the effect of the individual aerials will be additive. An increase of directivity in the horizontal plane can be achieved by arranging additional bays of

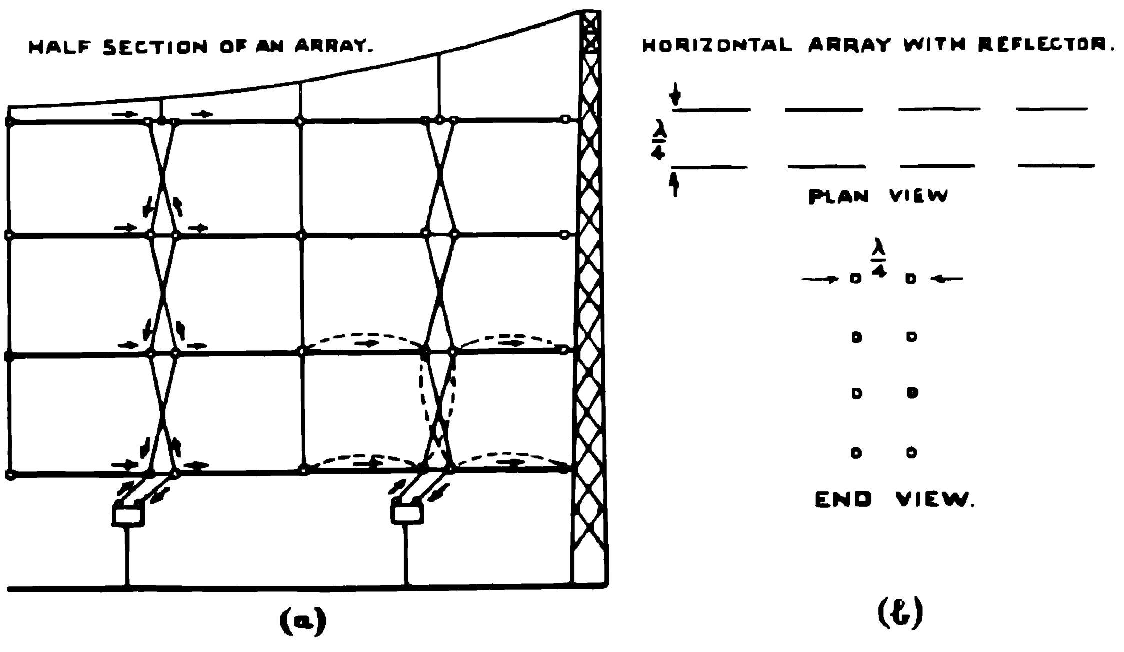


FIG. 51.

stacked aerials, *i.e.*, by arranging to have co-linear array elements. The horizontal polar diagram will be much the same as that of a broadside array of vertical aerials ; it may be made uni-directional by arranging a similar array of reflector aerials at a distance of a quarter wavelength from the directly excited ones. The nett result is to produce a beam system with low angle radiation, without the use of phasing coils or high aerials.

Horizontal arrays are widely used and the half wavelength elements may be connected together in many different ways. Fig. 51 (*a*) shows two bays, each with four sets of **doublets,** each doublet arm being a half wavelength long. The four sets of doublets are fed in parallel from the same transmission line, but the transposition of the latter is necessary in order that the vertically stacked horizontal elements may oscillate in phase with each other. Fig. 51 (*b*) gives a plan and end view of an array with reflector.

This type of horizontal array is used in the Service in some shore stations engaged in working certain point-to-point services. More recently, it has been adopted by the B.B.C. for the Empire broadcasting services radiated from Daventry. For this special service, four bays, each of four horizontal half wavelength dipoles, have been found to give satisfactory results. It produces a beam having a divergence of 20° on either side of the centre line. If only two bays were used, the beam divergence would be 34° on either side of the centre line. Reflector aerials are not at present being used in connection with Empire broadcasting. The arrays radiate strongly in two opposite directions with the result that, for example, Canada's evening programme is India's breakfast entertainment. In this way directional transmission is carried out in a total of 13 directions.

52. **The Beverage Aerial, or Wave Antenna.**—This consists of a wire ranging from a half to about 10 wavelengths long, supported horizontally at 10 to 20 ft. above the ground like a telephone

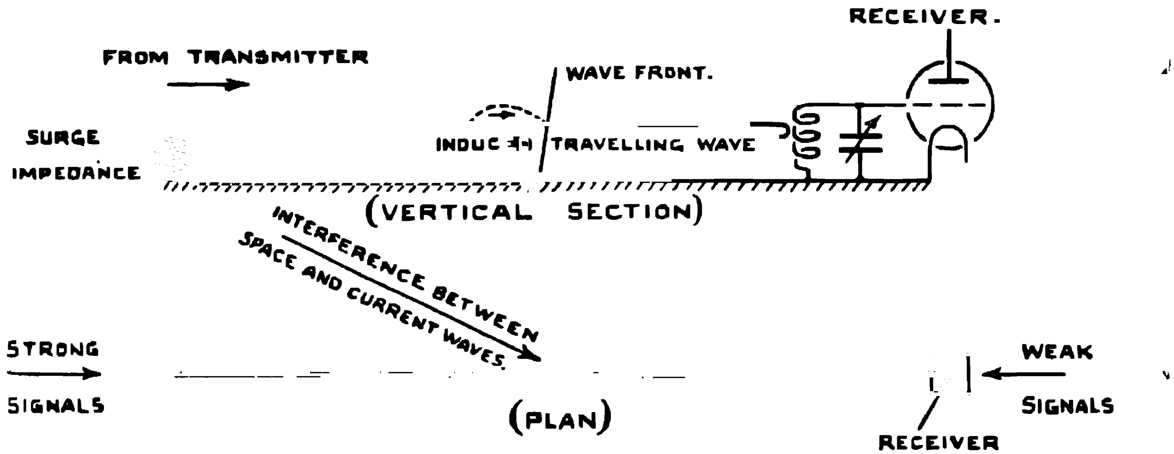


FIG. 52.

wire on the roadside, and running in the direction of the great circle bearing of the transmitter, as shown in Fig. 52. The end nearer to the transmitter is terminated to earth by an impedance equal to the surge impedance of the line, and the other end is joined through a matched impedance coupling to the input of a receiver.

This aerial was originally developed by H. H. Beverage as a long wave directional aerial, to be used for reception. In principle, its action at low frequencies depends upon the tilting of the wave front of the direct ray, due to absorption of energy by the surface over which the wave passes (*Cf.* paragraph 16). With reference to Fig. 16, a field of strength X will then have a horizontal component X sin θ, and will develop a voltage of that amount per unit length of wire. A travelling wave is initiated in the wire, and, if its velocity of propagation is assumed to be the same as that of a wave in free space, it will continue to receive energy as it travels and its amplitude will be built up

Thus, if the end of the aerial remote from the transmitter is directly connected to the grid of a valve whose filament is earthed, an oscillatory P.D. is developed between grid and filament.

Since the distributed impedance alters abruptly at the valve terminals, the travelling wave is therefore reflected and travels backwards to the open end of the aerial where again it suffers reflection. Stationary waves will be set up, giving rise to large aerial losses and a diminished P.D. at the valve. To prevent this, the long wire system must be made non-reflective, *i.e.*, non-resonant, by matching its surge impedance at both the receiver and the ' open end '—or end nearer the transmitter. The impedance at the open end also serves to absorb any travelling waves which are initially generated in a direction towards the transmitter.

The strongest signal will be received from a transmitter located in a direction along which the long wire points. A signal arriving from the opposite direction will initiate a wave travelling in a direction away from the receiver ; its energy will be absorbed by the terminating resistance, and little or no effect will be produced at the receiver In the case of a transmitter located in a direction at right angles to the length of the aerial, equal in-phase E.M.Fs. will be produced at points which are at different distances from the receiver. Partial cancellation of their effects will result, and the strength of the received signal will be relatively weak.

The directional characteristics partly depend upon the length of the aerial which, in the case of L/F, may be as much as 5 miles ; it also depends upon the terminating impedance and the velocity of propagation of the wave along the wire. By careful adjustment of the terminating impedance, it is possible entirely to cancel out any back-end reception ; misadjustment produces cancellation in other directions. Horizontal directivity may be increased by using an array of Beverage aerials connected in parallel. It is an extremely efficient directional aerial, inexpensive to construct, and, being untuned, is capable of giving simultaneous or successive reception on a number of different frequencies from transmitters in the same direction. Since its action as a receiver depends upon the presence of earth losses, it cannot be used satisfactorily as a transmitter.

In the Service, a simple wave antenna is used sometimes as an alternative aerial for H/F reception at some shore stations engaged in working certain point-to-point services ; usually its length is from 500 to 1,000 ft. In this case we are concerned with the reception of the indirect ray, and the energising of the aerial depends upon the horizontally polarised component of the wave. In its simple form it is not so efficient at H/F as it is at L/F ; for use at H/F its simple form has been adapted for higher efficiency by methods which will not be described here. The H/F " Beverage " aerial may also be used for transmitting.

53. **The Bruce Inverted " V " and Horizontal Diamond Aerials.**—In 1931, E. Bruce, of the Bell Telephone Laboratories, described two relatively simple H/F aerials, each combining the requisite directivity with satisfactory response over a wide frequency range without adjustment.

The use of **tilted wire aerials** may be regarded as the logical development of the Beverage type. If a wave front is not tilted downwards, or if there is no horizontally polarised component, no result will be produced in a horizontal Beverage aerial ; it must be tilted up to meet the wave, and for best results the angle of tilt is quite a definite one. The mechanism of the process is not quite the same as for a Beverage aerial. It can be most readily understood by adopting Bruce's vectorial method of explanation, considering first a half wave vertical aerial.

Fig. 53 (*a*) represents a vertical half wave aerial earthed through a load impedance. A vertically polarised field will not produce an E.M.F. in the aerial in a manner already well understood ; however, for use in other not so simple cases, it is convenient to regard the nett effect as equivalent to the sum of a number of small causes at different distances from E. Each elementary length will have an induced voltage in it, and by considering the various small E.M.Fs. to be concentrated at points A, B, C, D and E, it is clear that in this case they will all be equal and in phase, as shown in Fig. 53 (*b*). Assuming the velocity of propagation of a wave disturbance in a wire to be that of light, a finite time will be taken before a " cause " at A produces an " effect " at E, and, in the case of a wave, by the time it arrives it will be lagging by 180° on a disturbance starting at E, since the distance

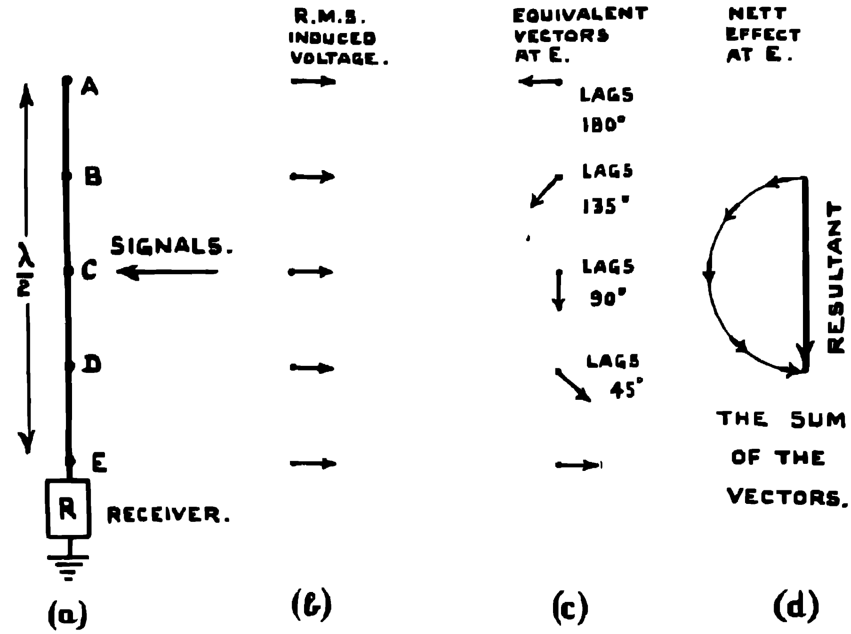


FIG. 53.

is a half wavelength. The waves from intermediate points will arrive at E lagging by intermediate angles, and the equivalent vectors to be added are shown in Fig. 53 (*c*). The process of vector addition produces the vector polygon of Fig. 53 (*d*), in which the vectors may be taken almost to lie along the arc of a semi-circle, the resultant being the closing side ; since that is the diameter, it is

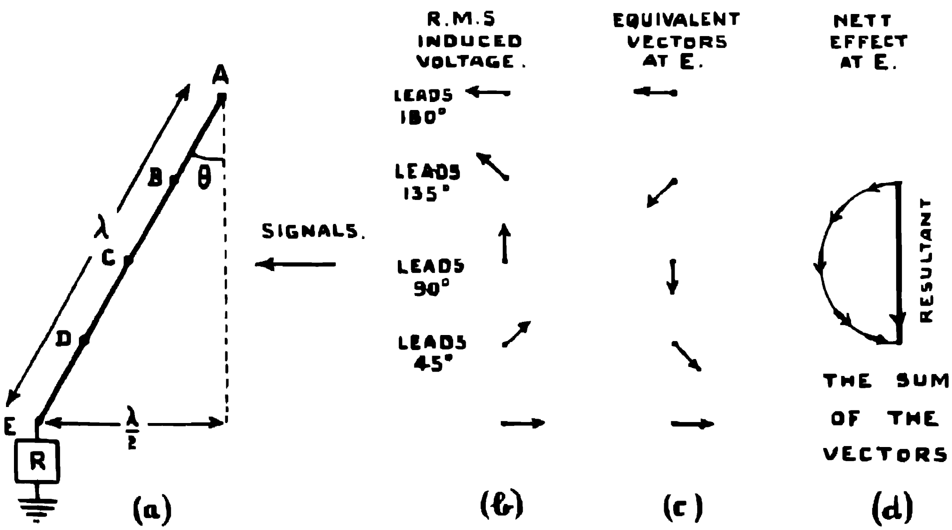


FIG. 54.

clear that these conditions produce a maximum E.M.F. in the aerial. It will be noted that the E.M Fs. in the elementary lengths at A and E are 180° out of phase ; it will be shown later that this may be adopted as the basis of the criterion by which the optimum conditions may be reeognised.

It will be noted that if the wire were vertical and one wavelength long, the vector sum would be given by a closed circular figure, implying the absence of a resultant. With reference to Fig. 43, it will be recalled that a wavelength aerial produces no field in the horizontal plane through the foot of the aerial This case may be compared with that of a horizontal Beverage aerial when the transmitting station is located in a direction at right angles to that along which the aerial points.

The null effect due to the lagging waves from the upper end of a wavelength aerial may be countered by tilting the aerial forward in the direction of the transmitter, as shown in Fig. 54 (*a*). With a plane vertical wave front, when the projection of the aerial on the horizontal plane is a half wavelength, the induced voltage vector at A will be 180° ahead of that at E, a point further from the transmitter and therefore lagging in phase. (*Cf.* Section " T," paragraph 4.)

Fig. 54 (*b*) shows the relative phase of the R.M.S. induced voltage vectors at intermediate points at the same instant of time. As before, the nett effect at E will depend upon how far the induced wave has to travel in the wire from each point. Fig 54 (*c*) represents the equivalent vectors to be added at R. Starting at E, the vector *e* represents the effect. At D, the initial E.M.F. is 45° leading, but since its distance in the wire from E is $\lambda/4$, its equivalent effect is that of vector *d* which lags by 90° on the initial cause. Similarly at C, the initial E.M.F. is 90° leading but since its distance from E is $\lambda/2$, the equivalent vector must be lagging by two right angles on the initial cause ; it is therefore vector *c*. (These vectors are not labelled in the Fig.)

Fig. 54 (*d*) represents the sum of these equivalent vectors at E. Again, the vector polygon is a semi-circle and the nett E M.F. in the tilted wire is a maximum. It is apparent that the criterion for optimum conditions already stated, still applies here, but in terms of the tilt angle, it is now better stated as follows :—

The E M.F. will be a maximum when the tilt is such that the length of the wire is a halfwave length longer than its projection upon the horizontal plane in the wave direction of propagation.

To get a larger effective induced voltage, the tilted wire must be made long with respect to the wavelength in use, and for each length of wire it is easy to calculate the proper tilt angle. It is here given in Fig. 55. It will be noted that when the length of the aerial is large, the tilt angle changes slowly, a fact which implies a wide frequency response on either side of the optimum frequency ; in practice it may safely range from 0·7 to 3·0 times the fundamental wavelength.

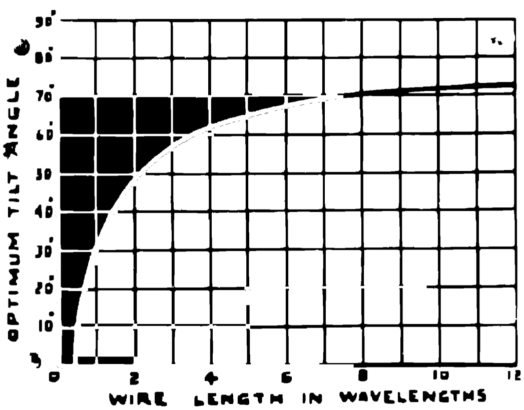


FIG. 55.

Since the tilt angle will only be correct for one direction of advancing waves, it is clear that the aerial has directional properties, and will discriminate between signals arriving from various directions.

Fig. 56 represents the addition of a second tilted wire, forming an inverted " V." It is shown earthed through a resistance R_1 equal to the surge impedance of the system. The earth, as usual, functions as an image aerial.

In the wire tilted towards the transmitter, the phasing of the induced voltage vectors LAGS progressively more from A to E ; in the wire tilted away from the transmitter, it LEADS from A_1 to E_1. Considering the two wires separately, this means that the nett effect at A_1 due to the one is the same as that at E due to the other. Jointly, the two elements of the " V " add in proper phase relation, a fact which is brought out by rotating through 180° each of the vectors representing the induced volts at points A, B, etc. The results of this process are shown in Fig. 56.

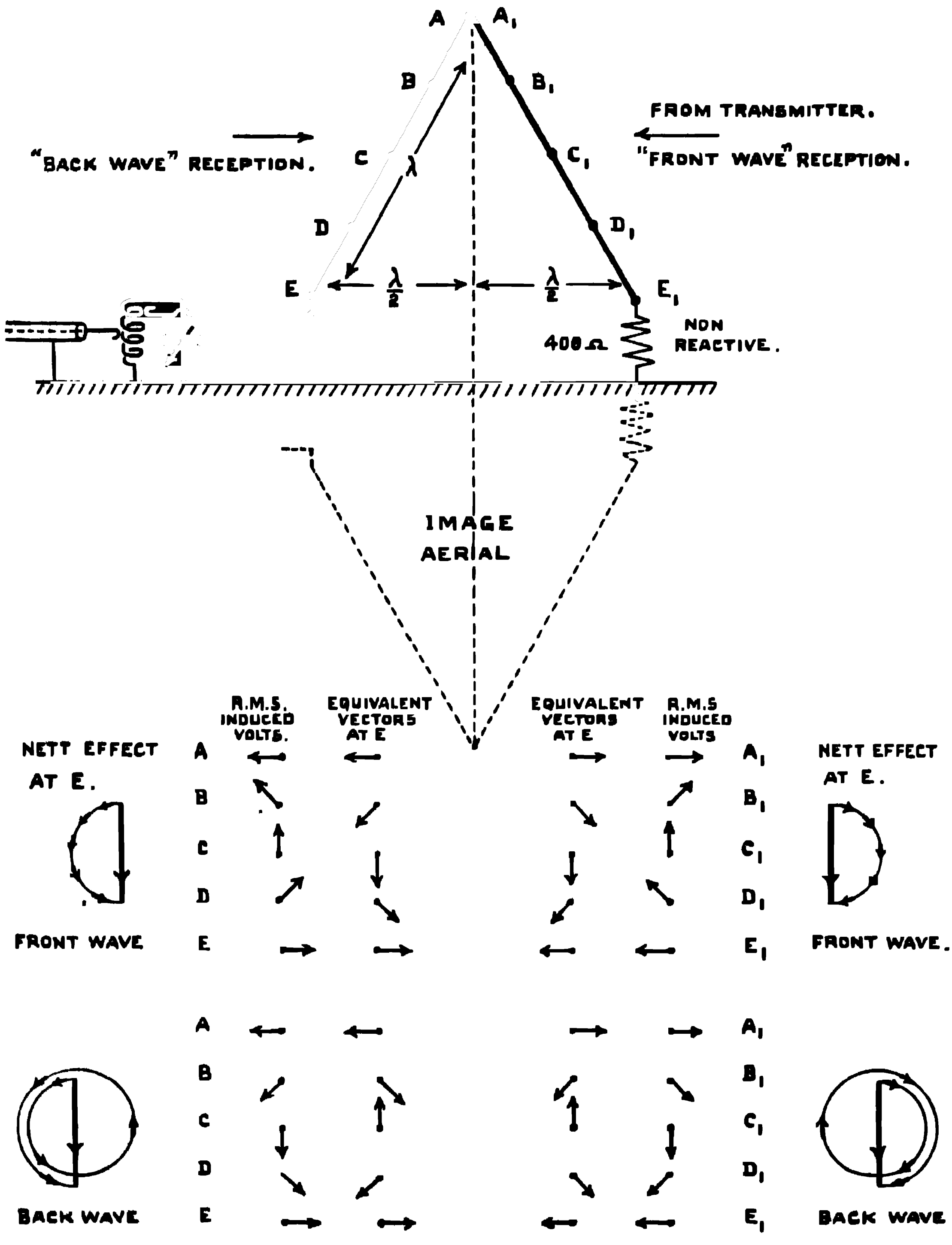

A
A1
B
B1
C
C1
D
D1
E
E1
λ
λ/2
λ/2
FROM TRANSMITTER.
"BACK WAVE" RECEPTION.
"FRONT WAVE" RECEPTION.
400 Ω
NON REACTIVE.
IMAGE AERIAL
R.M.S. INDUCED VOLTS.
EQUIVALENT VECTORS AT E
EQUIVALENT VECTORS AT E
R.M.S INDUCED VOLTS
NETT EFFECT AT E.
NETT EFFECT AT E.
FRONT WAVE
FRONT WAVE.
BACK WAVE
BACK WAVE

FIG. 56.

In addition to giving an increase in signal strength, the inverted " V " aerial presents a construction for which the non-reflective termination R_1 is more easily arranged than for a single ..ted wire. Moreover, it has enhanced directional properties, and, if desirable, may be arranged so that " back wave " reception is entirely cancelled. In front wave reception, principal travelling waves are set up in the direction of the receiver ; any in the other direction are absorbed in R_1. In back wave reception, the principal travelling waves will proceed towards R_1 where their energy is absorbed ; the effect of any energy travelling towards R may be summed up by reference to Fig. 56, where it will be observed that the phase of the induced voltage at (say) C lags behind that at E. Moreover, since its distance from E is a half wavelength, its effect at E makes it equivalent to a wave lagging by an additional 180°.

It will be seen that in this case the phase changes rapidly, and the appropriate vectors are shown in Fig. 56. The elementary vectors trace 1½ circles, and with a longer wire would trace more. Some residual effect is therefore produced at R ; in fact, although the aerial is correctly designed for maximum E.M.F. in the case of front wave reception, it represents the worst case if it is hoped to eliminate back wave reception. Bruce shows that it is possible simultaneously to satisfy these two requirements by designing the aerial so that the wire length of each element should be an odd number of quarter wavelengths greater than one, while at the same time satisfying the condition for optimum tilt, already stated. Where so designed, the vector diagrams for back wave reception will be closed figures.

There are various alternative forms of these aerials. Instead of terminating to earth, and relying on the latter to produce the effect of an image aerial, a double " V " or **vertical diamond aerial** may be set up and the whole insulated from earth. The lower half replaces the earth in its functions, and the aerial responds well to vertically polarised waves. The " **horizontal diamond** " aerial will respond to horizontally polarised waves, and bears the same relation to the inverted " V " as a horizontal dipole does to a Marconi aerial. The horizontal form frequently gives better results in districts where there is much static interference. For matching purposes, their impedance is of the order of 600 ohms ; the actual figure may be controlled by constructing the legs of the diamond with spaced wires.

The horizontal diamond type has vertical plane directivity which is controlled by the length of the legs, the tilt angle, and the height above the ground. As usually operated it possesses low angle vertical directivity. In general, both the vertical and horizontal polar diagrams resemble those of a broadside array.

The simplicity, satisfactory performance, and low constructional cost of these aerials has done much to make them almost more popular than complicated directional arrays. In the Service they find some use at certain shore stations engaged in working point-to-point services. The B.B.C. have used them especially in connection with the reception of American broadcasting ; in that case an inverted " V " was employed in conjunction with a horizontal diamond, using a species of diversity reception as a means of countering the fading which is always experienced on H/F. It seems possible that they may find a limited application on board ship ; an inverted " V " type has been used in R.M.S. " Queen Mary." For the reception of the Empire broadcast transmissions, the B.B.C. have recently recommended their overseas listeners to use an inverted " V " aerial.

54. **Diversity Reception.**—This is the name given to the various schemes that may be utilised at a receiving station, the object of which is the elimination of the phenomenon of fading. In paragraph 34 brief reference was made to the cause of the latter ; in general, it may be said to be due to an interference effect produced by waves arriving at the receiver, having followed different paths.

Fading is more serious at high frequencies, where a given change in length of path constitutes a larger percentage of a wavelength ; it must be overcome if full use is to be made of the advantages of H/F transmission. It is usual for broadcast frequencies only to suffer badly from fading during night time reception, when the proportion of indirect ray is generally greater.

At the *transmitting end*, fading is countered by aerial design (*cf.* paragraph 40) and by the use

of interrupted continuous waves at H/F. In order that the latter should work most effectively in conjunction with the arrangements at the receiving end (**diversity reception**), the two-side band frequencies should not be far apart ; this involves the use of low modulating frequencies (from 50 to 300 cycles, depending on the wave frequency), and the best results are obtained by employing a carrier suppression system. I.C.W. transmission using modulating frequencies of the order of 1,000 kc./s. is of little help in overcoming fading, since it involves too great a separation between the aerials used for diversity reception ; it affords no help at all to a receiver on board ship employing a single aerial.

At the *receiving end* fading may be countered by methods based on the observation that it is a comparatively local phenomenon ; if a signal fades at one place, it may be quite strong at another a few hundred feet away. On this assumption two or more spaced aerials or arrays may be set up, and arrangements may be made to obtain a common output, so giving what is known as "**diversity reception.**" Including telegraphy and telephony under the same heading, there appears to be at least three ways of achieving this object, classified as follows :—

(*a*) MULTIPLE DIVERSITY.—A complete receiver is fitted to each aerial or array, and the A/F output is combined across a common impedance at the final output stage. It is necessary to use separate receivers for the aerials in order to avoid the possibility of anti-phase conditions at a common input. This is a very much used

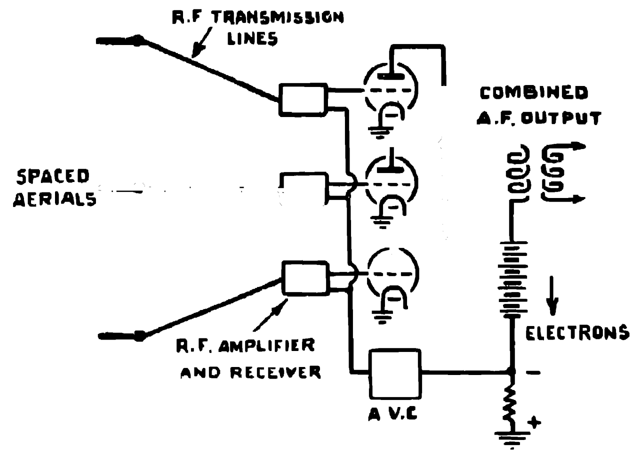


FIG. 57.

method, but it suffers from the disadvantage that if a deep fade occurs at one aerial, though it then contributes no signal, the noise output remains the same. This will produce an increase in the signal/noise ratio. Furthermore, if any aerial receives a distorted signal, due to selective fading, this distortion is handed on through the receiver.

(*b*) MULTIPLE DIVERSITY WITH COMMON A.V.C.—Each aerial is provided with its own receiver, but the final output is chosen automatically from that circuit which is giving the best results. This is achieved by using a common A.V.C. voltage for each receiver, provided in the manner indicated in Fig. 57. In effect, the receiver experiencing the strongest signal provides the A.V.C. voltage which reduces the gain of the R/F and I/F amplifying stages. The noise contribution of the "idle" aerials is reduced, and, with square law detection, the voltage output of the "active" aerials is relatively increased. This means that the receiver having a signal (say) three times the strength of another will contribute about nine times the output. This method is to be regarded as an improvement on simple "multiple diversity."

(c) SWITCHED DIVERSITY.—This refers to the use of two spaced aerials with a common receiver. The possibility of anti-phase conditions existing at the input is overcome by arranging automatically to switch the input between the two aerials at an audio-frequency of about 500 cycles per second. In this way, if the signal from aerial A falls to zero, it is still possible to rely on aerial B, although it is clear that the latter will only be connected to the receiver for 50 per cent. of the time. The switching cannot be done mechanically, and is usually done by controlling the grid bias of the valves at the inputs of the two receivers. A local oscillator coupled to the grid bias circuit may give an injected voltage which could be used to provide the equivalent of an electrical switch operated at a frequency of 500 cycles per second.

"Switched diversity" is suitable when the signal/noise ratio is high; since the principle of using one aerial for only 50 per cent. of the time involves certain losses, at times when the signal/noise ratio falls, it is usually customary to revert to "multiple diversity."

The best spacing between the aerials has been investigated by various workers. It has been observed that the requisite distance between aerials expressed in terms of λ increases as the frequency increases, a fact which implies that the actual spacing remains more or less the same throughout a wide frequency band. Very roughly, the best spacing varies from 10 to 20 wavelengths, with a minimum of 4 wavelengths.

A form of diversity reception employing an inverted "V" and a horizontal diamond aerial has been referred to in paragraph 53. It is based on the fact that the vertically and horizontally polarised components seldom fade simultaneously by the same amount

EXAMINATION QUESTIONS ON AERIALS.

Numerical Examples.

1. If 100 kW. of energy are radiated from an antenna of 100 metres effective height at a frequency of 60 kc./s., what would be the strength of the electric field in micro volts per metre at a distance of 100 kilometres, assuming that no absorption effects are present?

   (Result: 3×10^4 μV/m. C. & G. Final, 1932)

2. The current measured at the base of a transmitting antenna is 220 amperes. The antenna is of inverted "L" form, with a radiation height of 160 ft., and the frequency radiated is 37·5 kc./s. What is the power radiated?

   (Result: 2·853 kW. I.E.E., October, 1926.)

3. An aerial has an effective height of 100 metres, and the current at the base is 450 amps. (R.M.S.) at 40 kc./s. What is the power radiated? If the total resistance of the aerial circuit is 1·12 ohms, what is the efficiency of the aerial?

   (Result: 57 kW.; 25·14 per cent. C. & G. Final, 1930.)

4. A transmitting aerial having an effective height of 200 ft., takes a current of 50 amps. (R.M.S.) at a frequency of 480 kc./s. Calculate :—

(a) The radiation resistance of the aerial.
(b) The power radiated.
(c) The aerial efficiency for a total aerial resistance of 50 ohms.

(Result : 15·09 ohms ; 37·73 kW. ; 30·17 per cent. L.U., 1933.)

5. A transmitting aerial having an effective height of 120 ft. takes a current of 10 amps. at 1,000 kc./s. If a receiving aerial with an effective height of 50 ft. and an effective resistance of 4 ohms is erected at a distance of 50 miles from the transmitter, estimate the current in the receiving aerial. Explain the formula and each step in the argument. What complications, if any, would be introduced into the calculation if the distance between transmitting and receiving aerials were 1,000 miles ?

(Result : 21·76 mA. L.U., 1934.)

Mainly Descriptive Examples.

1. What are the quantities that form the total "aerial resistance" ? Draw a curve showing how the various parts and the total depend on the wavelength.

(C. & G.I., 1927.)

2. Describe a method whereby the resistance of a large transmitting antenna can be accurately determined.

(I.E.E., May, 1933.)

3. State the various types of masts and towers used to support the antennæ of land stations. Discuss the relative advantage of stayed masts versus self-supporting towers for various heights and structures.

(C. & G. Final, 1931.)

4. Describe any type of directional antenna array suitable for short wave transmission or reception, and indicate clearly how it is connected to the feeders.

An antenna array consists of eight vertical aerials in a straight line spaced half a wavelength apart and energised equally in phase. What will be the angular width of the forward beam in the horizontal plane ?

(Result : 29°. C. & G. Final, 1935.)

5. Explain the causes of fading in the reception of short and medium waves. What methods are adopted in practice to minimise the effects of fading ?

(C. & G. Final, 1933.)

DIRECTION FINDING.

1. This section gives a treatment of the principles of methods which can be used to determine the direction of incoming radio signals. Assuming that these have travelled along a great circle path, as is normally the case, the bearing so determined will give the direction of the transmitter with respect to the receiver. This process is known as Direction Finding (D/F).

In general it can be done in two ways :—

(*a*) With "all round" transmission, which is the normal method in a ship, the bearing of its signals must be determined by the use of special apparatus at the receiving station. This is **directional reception**, and it is with this that the section is mainly concerned.

(*b*) Alternatively, as a result of previous information it may be known that at a certain time a transmitter is radiating energy on one definite bearing only. The reception of the signal by an ordinary receiver then gives the bearing of the transmitter. This is known as **directional transmission**, and is the principle of operation of the rotating navigational beacons used for the assistance of ships and aircraft. A brief discussion of this matter is included.

DIRECTIONAL RECEPTION.

2. **Basic Principles.**—The basic principle of directional reception is the use of a loop aerial or its electrical equivalent.

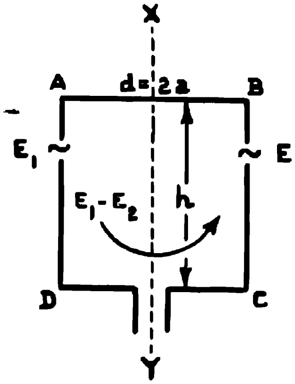


FIG. 1.

Fig. 1 shows this as a single loop of wire which for convenience only is taken as being rectangular in shape. It is to be considered as set up with its plane vertical and capable of rotation about a vertical axis XY. It will be recalled that for a vertical aerial of height "h" in a field of electric intensity X, a potential difference is produced between its ends given by $V = Xh$ in R.M.S. values, where the field strength X is usually measured in millivolts/metre. AD and BC function as vertical aerials, and if the plane of the loop is set at right angles to the path of vertically polarised radiation, it is clear that in all practical cases the variable electric field threading AD will be equal in magnitude and phase to that at BC at every instant, since both AD and BC may be regarded as equidistant from the transmitter.

This is illustrated in Fig. 2 (*a*). The variable electrostatic field threading BC is shown as being equal to that threading AD. E.M.F.'s represented by the vectors E_1 and E_2 are induced in sides AD and BC respectively. It is clear that in this case these E.M.F.'s are directed in opposite directions around the loop and, accordingly, cancel each other. It is at once evident that the horizontal sides can be neglected for a wave whose electric field is vertical, since the P.Ds. in these are negligible, and are in any case produced at right angles to the direction of these sides and so do not contribute to the E.M.F. around the loop.

Alternatively, one may regard the loop as being cut by the horizontal magnetic flux which is complementary to the vertical electric flux, the two together constituting an electro-magnetic wave. (This subject is more fully explained in the section on **AERIALS**). Disregarding the thickness of the wire forming the loop, AB and DC cut no horizontal magnetic flux; E.M.F.'s are only produced in AD and BC. This will remain the case whatever the orientation of the loop

Consider now that the loop is turned through 90°, until its plane lies in the direction of the wave, as illustrated in Fig. 2 (*b*). The distances of the two vertical sides from the transmitter then differ by the width of the loop. The electric fields of the wave are therefore different in phase at the two sides, and produce E.M.Fs. differing in instantaneous value and proportional to EF and GH respectively. At any instant there is therefore an E.M.F. around the loop equal to the difference between the E.M.Fs. in its vertical sides. The R.M.S. value of the E.M.Fs. induced in the two vertical sides will clearly be the same numerically, although they will be out of phase. In Fig. 2 (*b*)

vectors E_1 and E_2 represent this, while the vector triangle shows the value of $E_1—E_2$, the loop E.M.F. In any position of the plane of the loop other than that of Fig. 2 (*a*), the vertical sides are at different distances from the transmitter and, accordingly, the vectors E_1 and E_2 are not in phase, and a resultant E.M.F. equal to the vector difference of E_1 and E_2 will be produced around the loop. It is clear that the loop E.M.F. is a maximum in the case of Fig. 2 (*b*).

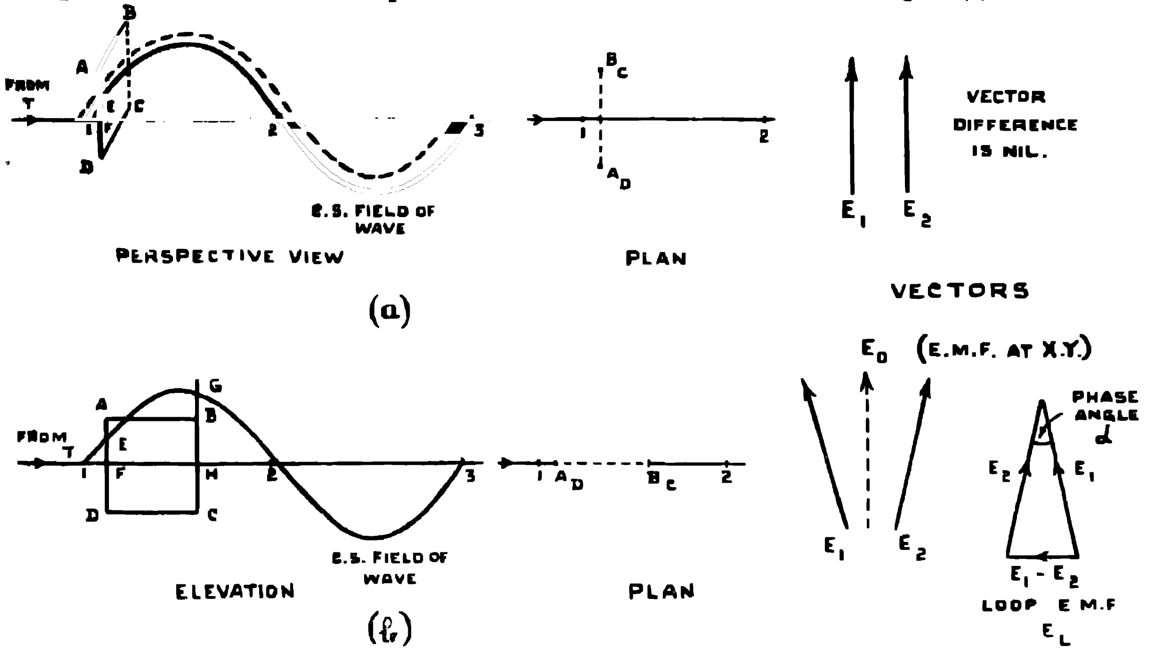


FIG 2.

If θ is the angle between the plane of the loop and the direction of the wave then writing E_L for the maximum R.M.S. value of the loop E.M.F., and E_θ for the loop E.M.F. at any angle θ, we have

$$E_\theta = E_L \text{ when } \theta = 0, \text{ and } E_\theta = 0 \text{ when } \theta = 90^\circ.$$

From this it would appear that E_θ varies as cos θ, and hence we write

$$E_\theta = E_L \cos \theta. \quad \text{(1)}$$

It should now be evident that the numerical value of E_L, the maximum R.M.S. value of the loop E.M.F., is proportional to

(i) h, the height of the vertical sides,

(ii) d, the width of the loop. This controls the phase angle and gives a maximum value of $E_1 - E_2$, when $\alpha = 180^\circ$, that is to say when $d = \frac{\lambda}{2}$ (*see* Fig. 2 (*b*)),

(iii) X, the intensity of the electric field.

From Fig. 2 (*b*) it is also clear that E_L, will vary inversely with λ, the wavelength. Further, when *d* is small in comparison with λ, it follows that since it is advantageous to make both *h* and *d* large, this is really equivalent to saying that A, the area of the loop, should be as large as possible. It can be shown that the following formula relates these quantities together :—

$$E_L = \frac{2\pi XA}{\lambda}, \text{ and hence } E_L = \frac{2\pi XAf}{c}, \quad \text{(2)}$$

(where $c = f\lambda$, *c* is the velocity of propagation of E.M. waves, and *f* is the frequency).

In regard to the phase of E_L, it will be recalled that as for all vertical aerials, vectors E_1 and E_2 in Fig. 2 (*b*), representing the E.M.Fs. in AD and BC respectively, are **in phase with the field of the wave.** E_0 represents the relative direction of the field of the wave at the centre of the loop and it is clear from the geometry of the figure that E_0 is at right angles to $E_1 - E_2$. **This means that E_L, the loop E.M.F., is 90° out of phase with the field of the wave.** To understand the problem of sense finding it is important to bear this in mind.

The loop extracts energy from passing radio waves but is not a very effective aerial from the point of view of signal strength, because E.M.Fs. induced in the two vertical sides are in anti-phase and so tend to cancel each other. This can be expressed in terms of the effective height, which is the resultant loop E.M.F. divided by X, the field strength. This gives

$$\text{Effective ht.} = \frac{2\pi A f}{c}, \text{ or in terms of } h$$

$$\text{Effective ht.} = \left(\frac{2\pi f d h}{c}\right) \cos\theta,$$ for a rectangular frame at any angle θ to the direction of the signal.

The effective height of a simple quarter wave aerial is $2/\pi$ times its physical height (R10), so taking a frame and a simple aerial each of the same height, and using the frame aerial in the position of maximum signals, we obtain the ratio—

$$\frac{h_2}{h_1} = \frac{\text{Effective height of frame aerial}}{\text{Effective height of simple aerial}} = \frac{2\pi f d h/c}{2h/\pi} = \frac{\pi^2 f d}{c},$$

and for a frame of width 1 metre, at 600 kc./s , the value of c being assumed to be 3×10^8 metres/sec.

$$\frac{h_2}{h_1} = \frac{10 \times 6 \times 10^5}{3 \times 10^8} = \frac{1}{50}.$$

It is thus clear that sensitive pre-selector gear is required in D/F work, or when a frame aerial is used for ordinary reception.

3. **Polar Diagram of Reception.**—We have

$$E_\theta = E_L \cos\theta.$$

A curve may now be completed showing the magnitude and phase of E_θ for all values of θ from 0° to 360°. Such a curve is called a Polar Diagram of Reception, and in this particular case of the loop aerial it takes the form of two circles as shown in Fig. 3, and is usually called a " figure-of-eight " diagram.

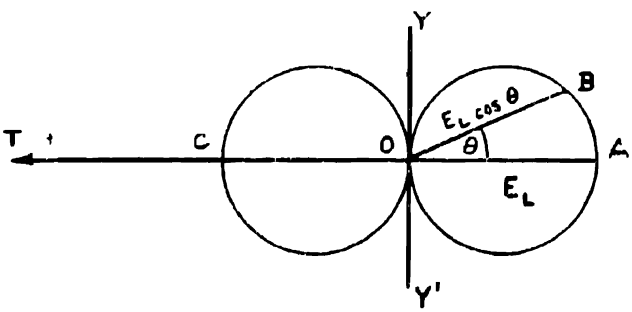


FIG. 3.

The value of E_θ for any particular value of θ is obtained by drawing through O to the curve a line making a counter clockwise angle θ with OA, *e.g.*, OB is E_θ for the value of θ shown in Fig. 3

$$OA = OC = E_L.$$

It can be seen that the curves are circles by considering that they have to satisfy the relation $E_\theta = E_L \cos\theta$. In the particular case shown, this means that $OB = OA\cos\theta$. Hence the angle OBA must be a right angle, and B is therefore a point on a circle whose diameter is OA.

That there must be two circles, arranged as shown, is evident from the fact that in any position the loop may be turned through 180° without affecting the numerical value of the E.M.F. induced round it ; the phase of the E.M.F. is, however, altered by 180°, *i.e.*, the E.M.F. is in the opposite direction round the loop at any corresponding instant compared with the first position. The polar curve can be taken as indicating the proportional strength of E.M.F. induced in a loop aerial which is rotated through 360°, the transmitting station being fixed ; or, alternatively, the E.M.F. induced in a fixed loop by a ship steaming round it in a circle and transmitting continuously. With the loop at right angles to TO, *i.e.*, along YOY', zero signals are received.

In the same way, the polar diagram of a vertical open aerial would be represented by a circle drawn with O as centre, indicating equal receptivity from all directions.

The variable E.M.F. induced in a loop aerial according to its orientation is of great assistance in direction-finding, and, in combination with the E.M.F. of a vertical aerial, for sense finding. It

is essential to bear in mind, however, that the vertical aerial E.M.F. is in phase with the electro magnetic wave, but that the loop aerial E.M.F. is 90° out of phase with the electromagnetic wave. **The elimination of this phase difference so that the effects may be capable of arithmetical addition or subtraction, is the main feature in the design of sense-finding receivers.**

★4. **Mathematical Analysis.**—To deduce the formula giving the numerical value of E_θ, the loop E.M.F., it is necessary to consider the actual value of the phase difference between the two vertical sides, when the plane of the loop makes an angle θ with the direction of the transmitter. Considering Fig. 2 (*b*), the electric field at F can be written in the form $\mathscr{E}\sin\omega t$, $\mathscr{E}$ being the amplitude or maximum value of the electric field. The magnetic field at the same point may be written in the form $\mathscr{H}\sin\omega t$, $\mathscr{H}$ being—likewise—the maximum value. Points 1 and 3 in this diagram are distant λ apart and differ in phase by 2π; by simple proportion points d apart will differ in phase by the angle $\frac{2\pi d}{\lambda}$. Further, since the whole wave system moves forward, with respect to any fixed point F,

points FURTHER from the transmitter than F, LAG in phase,
points NEARER to the transmitter than F, LEAD in phase.

This means that any event, such as the instant at which the fields pass through their zero values, happens at H after happening at F; at the point 1 it will happen before it happens at F, since the wave motion reaches there earlier. Hence, if the fields at F are represented by $\mathscr{E}'\sin\omega t$ and $\mathscr{H}'\sin\omega t$, the fields at H can be represented by $\mathscr{E}'\sin\left(\omega t-\frac{2\pi d}{\lambda}\right)$ and $\mathscr{H}'\sin\left(\omega t-\frac{2\pi d}{\lambda}\right)$. If the distance from 1 to F is represented by s, the fields at 1 would be $\mathscr{E}'\sin\left(\omega t+\frac{2\pi s}{\lambda}\right)$ and $\mathscr{H}'\sin\left(\omega t+\frac{2\pi s}{\lambda}\right)$.

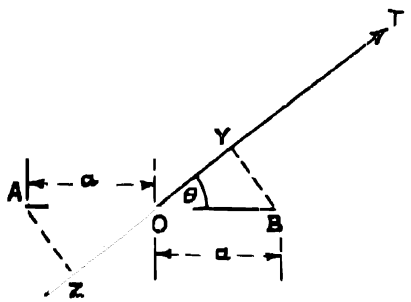


FIG. 4.

Fig. 4 shows the plane of the loop AB making an angle θ with the direction of the incident wave. AZ and BY are drawn perpendicular to the direction of the wave, and therefore lie in the wave fronts passing through A and B respectively. In other words, the field at A is equal to the field at Z and the field at B is equal to the field at Y. $AO = OB = a$, and $OZ = OY = a\cos\theta$. h = the length of the vertical sides of the loop.

Let the electric field of the wave at O be $\mathscr{E}'\sin\omega t$. Then the field at Z, at a distance $a\cos\theta$ *further* from the transmitter is $\mathscr{E}'\sin\left(\omega t-\frac{2\pi a\cos\theta}{\lambda}\right)$; similarly the field at Y is $\mathscr{E}'\sin\left(\omega t+\frac{2\pi a\cos\theta}{\lambda}\right)$.

The E.M.Fs. induced in the vertical sides AD and BC are therefore $\mathscr{E}'h\sin\left(\omega t-\frac{2\pi a\cos\theta}{\lambda}\right)$ and $\mathscr{E}'h\sin\left(\omega t+\frac{2\pi a\cos\theta}{\lambda}\right)$, and the loop E.M.F. is the difference of these two expressions. Hence

$$e_\theta = \mathscr{E}'h\left[\sin\left(\omega t+\frac{2\pi a\cos\theta}{\lambda}\right)-\sin\left(\omega t-\frac{2\pi a\cos\theta}{\lambda}\right)\right]$$

$$\therefore\quad e_\theta = 2\mathscr{E}'h\sin\left(\frac{2\pi a\cos\theta}{\lambda}\right).\ \mathrm{Cos}\,\omega t \quad\ldots\ldots\ldots\ldots \quad (3)$$

This is the general expression for the instantaneous E.M.F. round the loop. It is seen to be an alternating E.M.F., ($\cos\omega t$), of the same frequency $\left(\frac{\omega}{2\pi}\right)$ as the wave.

Since $\cos\omega t = \sin\left(\omega t+\frac{\pi}{2}\right)$, it follows that e_θ differs in phase by 90° from the field of the wave at the centre of the loop, *i.e.*, the mean field.

By putting $\theta = 0$, we get—

Amplitude of loop, E.M.F. $= 2\mathscr{E}'h\sin\frac{2\pi a}{\lambda}$, and when d, the width of the loop, is small in comparison with λ, $\frac{2\pi a}{\lambda}$ is a small angle, and the angle itself (in radian measure) may be substituted for its sine. The amplitude of the loop E.M.F. then becomes $2\mathscr{E}'h\frac{\pi d}{\lambda}$.

Now hd = A the area of the loop also $c = f\lambda$.

∴ Amplitude of loop E.M.F. $= \frac{2\pi\mathcal{X}'Af}{c}$, and in R.M.S. values we have $E_L = \frac{2\pi\mathcal{X}'Af}{c\sqrt{2}} = \frac{2\pi XAf}{c}$, which is the result that was stated in paragraph 2.

In general we have $E_\theta = E_L \cos\theta$.

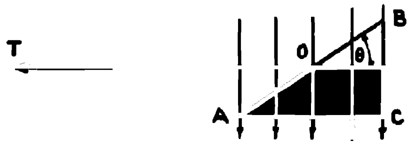


FIG. 5.

This value could also be deduced from the magnetic field, as in the case of a vertical aerial, by considering the flux cut by the two vertical sides. It is, however, perhaps more interesting to derive it from the magnetic field in terms of the flux-linkage with the loop.

The direction of the magnetic field is horizontal, and at right angles to the line TO in Fig. 5. Its value at O is $\mathcal{H}' \sin \omega t$. This may be taken as its average instantaneous value everywhere inside the loop, if the same assumption, that d is small compared with λ, is made.

The projected area of the loop in the direction of the transmitter, *i.e.*, at right angles to the magnetic field is

$$h \times d \cos\theta = A\cos\theta. \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots (Cf.\ \text{Fig } 5.)$$

Hence the flux-linkage with the loop is $\mathcal{H}'A\cos\theta\sin\omega t$.

The induced E.M.F. is the rate of change of the flux-linkage, or $\frac{d}{dt}(\mathcal{H}'A\cos\theta\sin\omega t) = \omega\mathcal{H}'A\cos\theta\cos\omega t$, in E.M.Us. if $\mathcal{H}'$ is in E.M.Us. Expressed in E.S.Us. $[\mathcal{X}' = c\mathcal{H}']$, it therefore becomes $\frac{\omega\mathcal{X}'A\cos\theta\cos\omega t}{c}$, the maximum amplitude is then $\frac{2\pi\mathcal{X}'Af}{c}$ as before.

5. The Frame Aerial.—The frame aerial is a development of the loop aerial, consisting of a number of loops instead of one.

If we could imagine n loops of the same size and co-planar, the E.M.F. induced in the frame would be n times that induced in a single loop. This follows, for example, from the proof given above that the E.M.F. is proportional to the rate of change of flux-linkage, the latter being n times as much for a loop of n turns as for a single turn. We have also the fact that the larger the frame aerial, the larger will be the E.M.F. induced. This latter effect is found in practice to be greater than the first, and it is usually better to have as large a frame as possible with a few turns, rather than a small frame with a large number of turns.

In practice the theoretical assumptions about frame aerials cannot be achieved, and if more than one loop is desired, so as to increase signal strength, the loops must be wound in either (*a*) box form, or (*b*) pancake form, as illustrated in Fig. 6. In the box form, the loops are of the same dimensions, but not coplanar, and are equivalent to a single loop of n times the area in a plane parallel to themselves, plus a loop at right angles of area equal to half the area of the vertical side of the box frame. Hence zero signals will not be obtained when the frame is exactly at right angles to the line joining it to the transmitter.

BOX FORM (a) PANCAKE FORM (b)

FRAME COIL AERIALS

FIG. 6.

In the pancake form the total E.M.F. is the sum of the separate E.M.Fs. in the loops, these being proportional to the dimensions in each case. It is equivalent to one loop whose area is the sum of the individual areas, and gives zero signals when its plane is at right angles to the transmitter.

In the case of the box form, with N turns, the formula of paragraph 2 now becomes

$$E_L = \frac{2\pi XNAf}{c},$$

and gives E_L in volts (R.M.S.) if all quantities are in practical and C.G.S. units. If we consider the magnetic field of the wave instead of its electrostatic field, then since $X = cH$, in absolute E.M. units, we have $E_L = 2\pi HNAf$; on dividing the result by 10^8, we have E_L in volts. In E.M.Us., and in air, H is expressed in "lines per square centimetre."

6. **D/F Systems.**—From the point of view of service practice, it is now possible to distinguish four systems of D/F and these will be treated in the order :—

(*a*) The rotating frame coil aerial system (paragraph 7).
(*b*) The crossed loop aerial or Bellini-Tosi system (paragraph 12).
(*c*) The fixed crossed frame coil aerial system (paragraph 16).
(*d*) The Adcock system (paragraph 38).

In general, the direction of the transmitter, as thus determined, is ambiguous. It is convenient, therefore, to make a distinction between the **direction** and **sense** of a bearing, the direction giving the line of the bearing and the sense indicating which of the two possible bearings is the correct one.

7. **The Rotating Frame Coil Direction Finder.**—To determine the direction of a transmitter use may be made of a vertical frame coil, or loop aerial, capable of rotation about a vertical axis. As the coil is rotated maximum signal strength is obtained when the coil points to the transmitter, and no signal is heard when its plane is at right angles to the direction of the transmitter. The determination, however, of either of these positions is sufficient to ascertain the direction of the transmitter. It will, however, be observed from the polar diagram, that the change in signal strength with the angle of rotation is much more pronounced in the neighbourhood of zero signals than in that of maximum signal strength. Thus, a more accurate determination of the bearing can be obtained from the zero signal position, and this position is always used in practice.

In regard to the design of the frame coil, for each frequency there is a best number of turns to give maximum efficiency. The lower the frequency the greater should be the number of turns. This arises from the fact that the E.M.F. induced in the frame increases directly as the number of turns, while the inductance increases approximately as the *square* of the number of turns. At the same time, the high frequency resistance increases more rapidly than the number of turns, depending considerably on the spacing of the turns and the diameter of the wire. The net result is that for any given small frequency range, there is an optimum number of turns for a coil tuned directly by a condenser. For example, for a coil 3 ft. 6 ins. in diameter six turns is a satisfactory number for a frequency band 500-1,000 kc./s. The coil can also be used for much lower frequencies without great loss of efficiency, but its use at higher frequencies is attended by an appreciable reduction in efficiency. At 15,000 kc./s., it would be preferable to use a single turn coil tuned directly by a condenser.

The coil may be rotated directly by hand operated mechanical means, or from a distant office by electric or hydraulic gear. Where the office is at a considerable distance from the coil, the capacity and inductance of the cables connecting the coil to the receiver will tend to affect the optimum number of turns for any desired frequency range.

The shaft which rotates the coil, directly or indirectly, may actuate a pointer moving over a scale graduated from 0 to 360°, and the pointer may be set so that the reading when zero signal is obtained is the true great circle bearing of the transmitter. If, for example, the wave is arriving from due north, the zero is obtained when the coil is pointing east and west, and the pointer may be set so that the reading is then 0°, *i.e.*, true north.

Coils in common use are of varied shape ; for some purposes, frame coils of hexagonal shape are to be preferred to those of square or circular form. As shown in paragraph 2, the area of the coil and *not* the shape is a factor of importance in determining the loop E.M.F.

8. **Antenna or Vertical Effect of a Loop.**—The use of a loop for direction finding depends entirely on the fact that when the loop is at right angles to the direction of propagation no signal is heard. Anything which may cause signals to be heard when the loop is in this position is detrimental to accurate direction finding. One such effect is due to the unequal action of the two sides of the loop as vertical aerials and is described as " vertical " or " antenna action."

It was seen in paragraph 2 that the E.M.F. round the loop was the difference between the E.M.Fs. induced in the two vertical sides. Now consider the simple D/F receiver shown in Fig. 7.

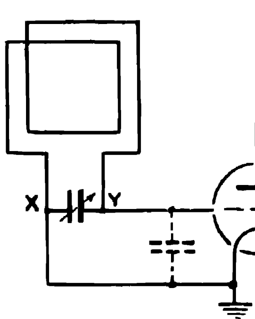


FIG. 7.

The loop (shown as consisting of two turns) is tuned by a condenser, and the voltage developed across the condenser is applied between the grid and filament of the first valve of a receiver. When the plane of the loop is at right angles to the direction of the transmitter, there is no resultant E.M.F. round the loop, and therefore no voltage across the condenser due to this cause. There are, however, large equal E.M.Fs. in the vertical sides of the loop. These vertical sides each have a certain amount of capacity to earth, one side from Y via the capacity to earth of the grid, and the other side from X via the filament. The sides of the loop therefore behave as earthed vertical aerials ; moreover, their paths to earth are unsymmetrical, and therefore of different impedance. It is obvious that the impedance from Y to earth via the grid will be different from the impedance from X to earth via the filament circuits. Thus, even when equal E.M.Fs. are induced in the two vertical sides (*i.e.*, in the position of the loop where zero signals would be expected), unequal currents will flow in them due to their unequal impedances, and the points X and Y will not be at the same potential. Hence a P.D. is developed across the condenser, and a signal will be heard. Due to this vertical effect, either no zero signal position will be obtained, or, if it is, it will not occur when the plane of the loop is at right angles to the direction of the transmitter.

9. **Elimination of Vertical Effect.**—The obvious method of preventing vertical effect due to unequal capacity of the vertical sides of the coil to earth, is either to equalise these capacities or to short-circuit them by providing symmetrical paths to earth of much smaller impedance for the vertical currents.

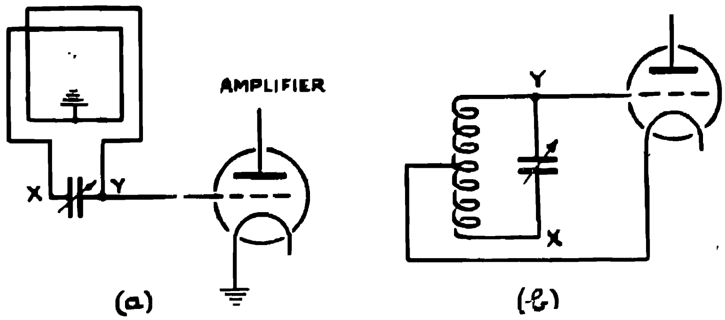


FIG. 8.

10. **Tuners and Receiving Apparatus.**—The difficulty of providing the requisite symmetry increases with the frequency, and is one of the chief features in the design of tuners or pre-selector gear in this and other systems of D/F. At low frequencies, below 100 kc./s. and up to 600 kc./s. for rough searching in the " stand-by " position, a simple solution is to earth directly the mid point of the coil as shown in Fig. 8 (*a*).

The vertical currents in the two sides of the loop then flow straight to earth, being provided with equal paths of much smaller impedances than those from the capacities to earth of the valve electrodes. Since the paths are symmetrical, X and Y will be at the same potential as far as vertical currents are concerned. It will be noted that in this circuit only half the voltage developed across the tuning condenser by the loop E.M.F. is applied between grid and filament of the valve. Fig. 8 (*b*) shows the equivalent circuit and represents the arrangement used sometimes at medium frequencies.

A normal frame is treated as aperiodic for low frequencies up to about 180 kc./s., and is coupled to a tuner by an auto-transformer coupling (Fig. 9).

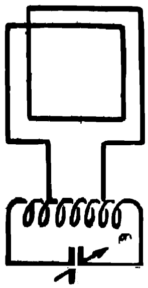


FIG. 9.

Between about 150 and 400 kc./s. it is the practice to tune the frame coil directly by a condenser, as in Fig. 8, the frame coil usually having an equivalent inductance of about 100 mics.

At higher medium frequencies from 400 to 600 kc./s., it is usually necessary to cut down the inductance of the loop by shunting it with an inductance of about 100 mics.

In each of these cases, for anything other than rough searching in the stand-by position, the input should be applied to a push-pull amplifier, to give the necessary selectivity and symmetry. Especially at the higher frequencies, the reactance of the paths to earth via the valve electrodes becomes comparable with that of the vertical sides themselves, and the only solution is to equalise these capacities.

A push-pull arrangement of valves readily lends itself to this purpose and a circuit suitable for medium frequencies is shown in Fig 10. For higher amplification, the triodes should be replaced by tetrodes.

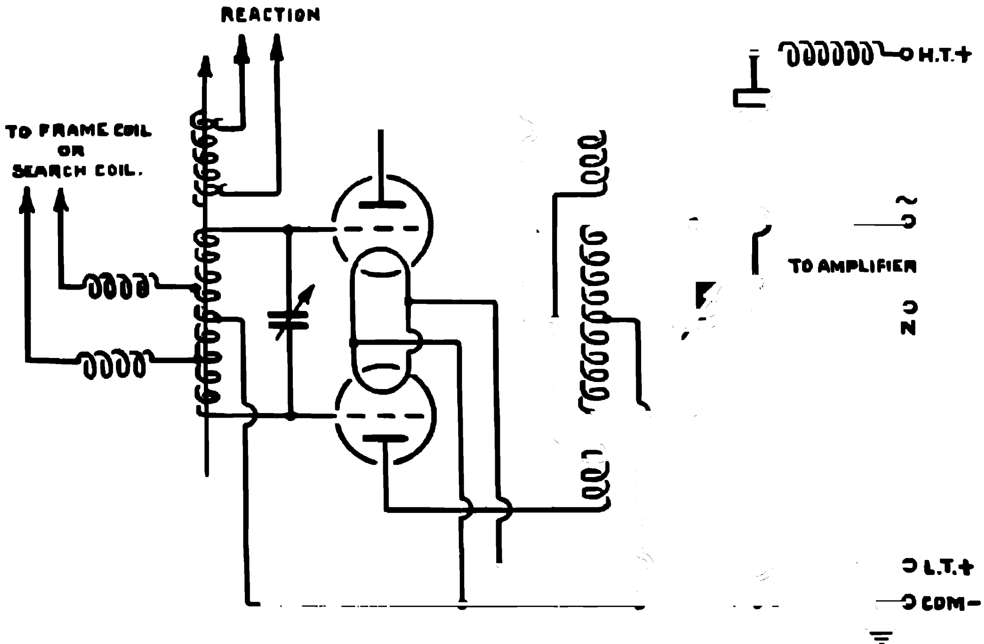


FIG. 10.

The reference to the search coil in Figs. 10 and 12 will be appreciated after paragraphs 12 and 14.

The "vertical" leads are then via the capacities to earth of the grids of the two matched valves, and so are of very nearly equal impedance. As far as vertical effect is concerned, the instantaneous potential of the grids of the two valves are, therefore, always equal. The corre-

sponding anode currents are also equal, and, flowing in opposite directions through the two halves of the output choke, give rise to zero resultant flux-linkage with the secondary. The loop E.M.F., of course, produces a P.D. across the tuning condenser, and therefore causes anti-phase potential variations on the two grids, giving a resultant E.M.F. in the secondary.

For frequencies above 700 kc./s. the frame is again treated as aperiodic, and is transformer coupled to a tuned circuit as shown in Fig. 11. The dotted line S indicates that there is an electrostatic earth screen between the windings. This prevents capacity action between the two coils which might introduce a lack of symmetry, but does not prevent the magnetic flux due to current in the aerial coil from linking with the coupling coil. On these frequencies the superheterodyne method of reception is used, in order to be able satisfactorily to amplify frequencies up to 15,000 kc./s.

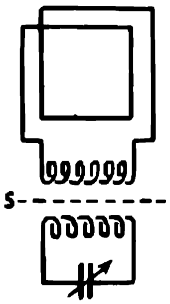


FIG 11.

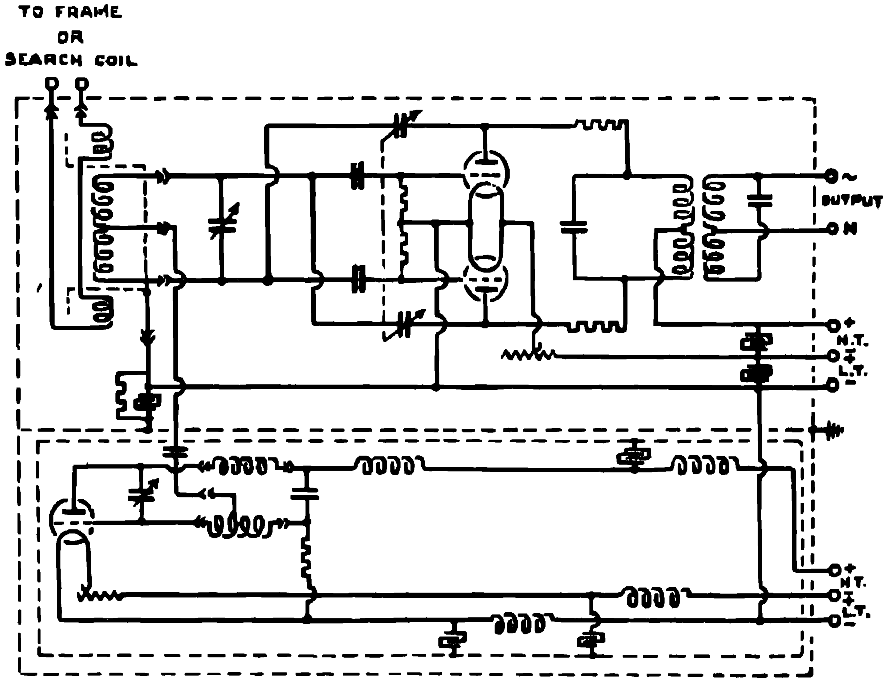


FIG. 12.

Fig. 12 illustrates a heterodyne detector unit, and Fig. 13 (*a*) is its simple equivalent circuit.

In Fig. 12 will be seen the special electrostatic screen fitted between the aerial coupling and the grid coupling coils. The magnetic coupling between these coils is usually arranged to be loose, so as to prevent the aerial circuit from causing excessive damping in the tuned grid circuit, should the aerial approach resonance at any frequency at which the model is being used. The ends of the grid coil are connected to the grid insulating condensers of two valves arranged in push-pull, which with the local heterodyne signal fed from the oscillator in the lower half of the diagram, functions as a beat rectifier. The anodes of the push-pull valves are connected through two R/F stopper resistances to a tuned anode circuit, resonant at a supersonic beat frequency of the order of 100 kc./s. The output is passed on to an amplifying stage the function of which may be considered to be similar to that of the intermediate frequency amplifier in a superheterodyne receiver.

The superheterodyne circuit is carefully screened and feeds through a small coupling condenser to the mid point of the grid coil; the general object of the design is to make the heterodyne circuit entirely independent of the signal frequency circuit which it is feeding, so that any alteration in the value of the tuning condenser will not affect the heterodyne circuit in any way. Since the filament circuits of the oscillator and push-pull combinations are common, it follows that the heterodyne input may be represented as in the simple equivalent circuit of Fig. 13 (*a*).

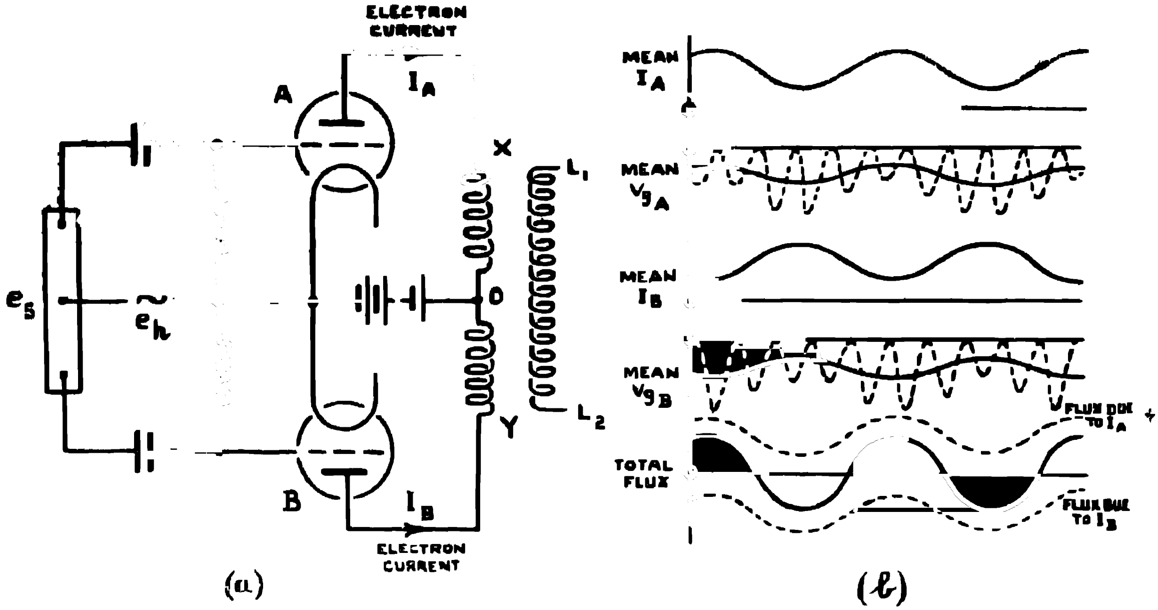


FIG. 13.

To understand the mechanism of beat rectification employing a push-pull circuit, it is essential to realise that :—

(*a*) The symmetry of such a system must be unbalanced by some means, since rectification always involves the action of an unsymmetrical or " one way " device.

(*b*) The valves must be operated as rectifiers.

An explanation of this circuit, in mathematical terms, is given in Section " N " (19). In this instance it may be simply explained by considering the following special cases :—

Case 1. (*Input* e_s). *only*.- In Fig. 13, an oscillatory input e_s will act differentially on A and B. Disregarding for the moment any grid current rectifying action, the instantaneous currents i_A and i_B will be in phase in XY and hence equivalent to one oscillatory current through the whole winding. Action in each half cycle is similar, and, accordingly, no rectification takes place. Due, however, to grid rectification at each grid, the mean anode current in each half of XY will be less than otherwise it would be. The combination will act as an ordinary push-pull amplifier.

Case 2. (*Input* e_h *only*).—An oscillatory input e_h applied in the position shown in Fig. 13 may be termed a " parallel " input. The grids at A and B will be operated always in phase, each being positive or negative at the same instant of time, and, accordingly, producing variations of anode current which are in anti-phase in XY and so produce flux changes which cancel out.

Case 3. (*Input* e_s *and* e_h).—In this case we will first assume that e_s and e_h are both of the same frequency and amplitude. We further assume that e_h is key controlled, that is to say, applied intermittently. Now if e_s and e_h are in phase at the grid of valve A, they will be in anti-phase at that of valve B. Moreover, if the amplitudes of each are equal, the effect at A is that of a signal of double the amplitude, while B continually experiences equal oscillatory voltages applied in anti-phase between grid and filament. In effect valve B is rendered inoperative, **the system is unbalanced,** and grid current rectification at A reduces the mean anode current for the period during which e_h is applied. If e_s and e_h are in anti-phase at the grid of valve A, the action will simply be reversed ; valve A will be rendered inoperative instead of valve B. If e_h has not the same amplitude as e_s, instead of rendering one valve or the other completely inoperative the effect will only be a partial one, and the decrease in mean anode current that represents rectification may not be as great. In place of grid current rectification one can substitute anode bend rectification with equal success. It is important to realise that, in some way, the two valves must be set to rectify,

for without this, even though one valve may be inoperative and the system unbalanced, no change in the value of the mean anode current will result. If e_s and e_h are not exactly in phase at either grid, or in anti-phase, it may still be possible to obtain a rectified output but the amplitude will not be as great.

Case 4. (*e_s and e_h not of the same frequency, and e_h not key controlled*).—If the difference in frequency between e_s and e_h is not too great, the phase difference between the two inputs at either grid will be continually changing. The nett effect is that the oscillatory input between grid and filament of either valve will be of a " beating " nature, and when the input to A is a maximum that at B will be a minimum ; this must be so, since when e_s and e_h are in phase at the grid of valve A, they are in anti-phase at that of valve B.

It must be remembered that the mean current I_A flows through the coil OX in the direction X to O and the mean current I_B flows through the coil OY from Y to O. The coils X and Y are wound in the same direction. Each of these currents produces a magnetic flux through the coil L_1 L_2 and these fluxes are in opposite directions. Now it is clear from Fig. 13 (*b*) that whenever the mean current I_A increases, the mean current I_B decreases and *vice-versa*, both of these currents varying at the beat frequency. Thus, whenever the magnetic flux through L_1 L_2 due to the mean current I_A increases, that due to the mean current I_B decreases and *vice-versa*. But since the main fluxes are in opposition the variations are additive, since an *increase* in flux in one direction through a coil produces the same effect as a *decrease* in flux in the opposite direction. This is illustrated in the bottom graph of Fig. 13 (*b*) where the two dotted curves show the individual magnetic fluxes produced in L_1 L_2 by the mean currents I_A and I_B, whilst the full line curve shows the resultant flux obtained by geometrical addition. This shows that the resultant in L_1 L_2 is a magnetic flux varying at the beat frequency. The total E.M.F. in L_1 L_2 thereby produced, is twice that which would be produced by coils OX or OY acting alone.

Finally, it will be observed the reaction condensers control the sensitivity of the receiver. In operating the outfit, care is taken not to increase reaction up to the point at which the circuit becomes self-oscillatory. It will be noted that filter chokes and decoupling condensers are fitted to prevent coupling between the superheterodyne circuit and the R/F circuit through the common battery leads.

11. **Screened Coils.**—Another method of reducing antenna action is to equalise the capacity to earth by surrounding the coil with a number of earthed conductors, which are not allowed, however, to form closed conducting paths. They may, for example, consist of a large number of vertical aerials immediately surrounding the coil ; but a simpler solution in practice is to enclose the windings of the coil in a metal tube, as shown in Fig. 14. Currents are prevented from circulating round the tube by the insulating segment shown at S, which is commonly made of porcelain or ebonite. The tube itself is earthed at the bottom at its centre point.

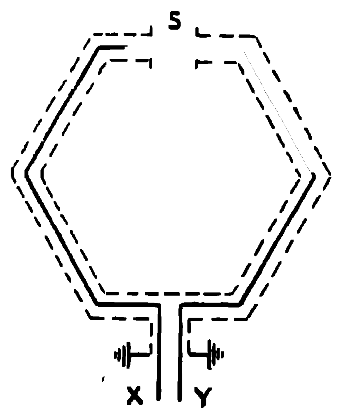


FIG. 14.

It is clear that comparatively small currents will flow up and down the screen ; for although the two vertical sides of the screen will certainly act as open aerials, they will be far from resonance. At all but very high frequencies, the impedance of each screen acting as an aerial will be very high, and the magnetic fields of the vertical currents in the screen very small. They really act as very inefficient vertical aerials. Thus the screen currents will have little effect on the total magnetic flux threading the circuit of the frame. This effect is entirely dependent on the segment S being a good insulator. If S is short-circuited, quite large currents will circulate round the screen and cancel the magnetic flux of the wave itself.

By Faraday's Law, the total E.M.F. acting round a closed conducting circuit is equal to the rate of change of flux-linkage through the circuit. As the amount of flux is scarcely influenced by the earthed screen, the E.M.F. induced in the loop is practically the same as if there were no screen. However the matter is regarded, it is certain that the earthed tube is **not a magnetic screen.**

The capacity between the sides of the loop and the earthed screen, however, provides much the greater proportion of the total capacity of the sides to earth, by virtue of which they behave as vertical aerials, and the valve electrode capacities to earth carry very little or none of the vertical current. Since the screen is symmetrical about the sides, the points X and Y are at the same potential, in so far as their potentials are a result of vertical effect.

It is common practice to use earthed screens for rotating coil direction finders, designed to work on the L/F and part of the M/F wave bands. Unscreened coils are usually used on the H/F band, since an earthed screen would tune at some particular frequency depending on its dimensions, and would then seriously reduce the efficiency of the aerial system.

12. **Crossed Loop Aerial or Bellini-Tosi Direction Finder.**—Instead of rotating the aerial itself it is possible to use a system of fixed aerials with an instrument called a radio-goniometer in the office. This avoids the mechanical or electro-mechanical difficulties which are involved when it is necessary to rotate a loop aerial. When a direction finder is installed in a ship this is often a matter of considerable difficulty. The principle will first be illustrated by a special case. (A goniometer is an "angle-measurer.")

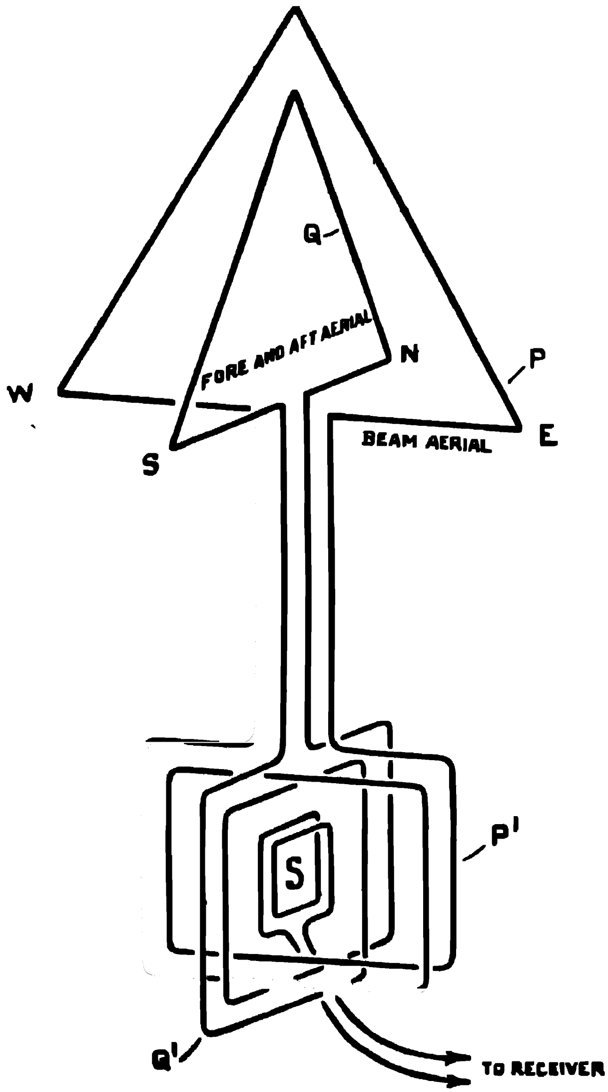


FIG. 15.

Referring to Fig. 15, suppose P to be a loop in a vertical plane pointing East and West, and Q a loop pointing North and South. P is joined by leads as shown to a coil P^1 inside the office. In Fig. 15, P and P^1 are shown in the same plane for simplicity, but this is not necessary in practice. In a similar manner Q is joined to a coil Q^1, which is mounted at right angles to the coil P^1. The coil S, which is joined to the receiver, can be rotated and is coupled magnetically to P^1 and Q^1. Consider a wave coming from a direction due North (or due South). No E.M.F. is induced in the loop P, and therefore no current flows in P^1; maximum E.M.F. is induced in Q, and a current flows in Q^1. If now the coil S is rotated it has maximum E.M.F. induced in it, and maximum signals are obtained when it is parallel to Q^1, *i.e.*, at right angles to the magnetic field of Q^1, and a zero is obtained when it is parallel to P^1. Similarly, for a wave incident from due East (or due West), a zero is obtained when S is parallel to Q^1. Again, suppose the wave to be incident on a bearing of 045°. Equal E.M.Fs. are then induced in P and Q, giving rise to equal currents in P^1 and Q^1. Zero signal thus occurs when S is inclined equally to P^1 and Q^1. Finally, we see that for a wave on a bearing of 000°, the position of S for a zero is 0° from the coil P^1; for a wave incident at 090° the position of S for a zero is 90° from P^1; and similarly, for a wave incident at 045°, the position of S for a zero is 45° from P^1. In fact, as the bearing of the wave changes, the direction of S for a zero changes by the same amount. This will be proved in the next paragraph for any angle of incidence of the wave.

★13. **Mathematical Analysis.**—The general case will now be considered. The two loops P and Q may be set up in any two vertical planes at right angles to each other. In order that each of them, treated as a separate loop aerial, may have the same E.M.F. induced in it by a wave passing over it in its own plane, both loops must have the same area, since the E.M.F. is proportional to the area. (It will be seen below that this may not be the case in practice, in order to correct certain errors that arise.) Suppose, now, that a wave whose electric field is $\mathscr{X}\sin\omega t$ is incident on the system at an angle θ with the plane of loop Q, and therefore at an angle $(90°-\theta)$ with the plane of loop P (Fig. 16).

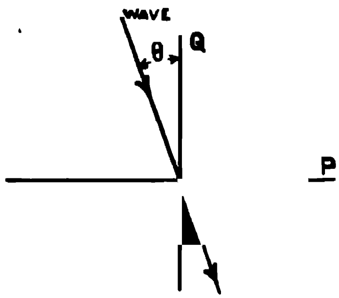


FIG. 16.

From paragraph 2 the E.M.F. in Q is $\frac{\omega\mathscr{X}A}{c}\cos\theta\cos\omega t$, and the E.M.F. in P is $\frac{\omega\mathscr{X}A}{c}\cos(90-\theta)\cos\omega t = \frac{\omega\mathscr{X}A}{c}\sin\theta\cos\omega t$, provided the widths of P and Q are small compared with the wavelength. The currents produced by these E.M.Fs. depend on the impedances of the two circuits, including their goniometer coils. In practice untuned aerials are used, and their resistance may be neglected, compared with their inductive reactance. If both circuits have the same inductance L currents proportional to the E.M.Fs. and lagging 90° in phase on the E.M.Fs. flow in the two goniometer windings P^1 and Q^1, the currents being given by $\frac{\omega\mathscr{X}A}{c\omega L}\cos\theta\sin\omega t = \frac{\mathscr{X}A}{cL}\cos\theta\sin\omega t$ in Q^1, and similarly $\frac{\mathscr{X}A}{cL}\sin\theta\sin\omega t$ in P^1.

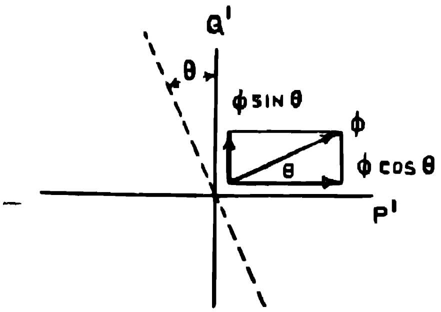


FIG. 17.

It may be noted in passing that these current amplitudes are independent of frequency. If the coils P^1 and Q^1 are identical, except that they are at right angles to each other, the fluxes produced by the currents bear the same proportion to the currents, and may therefore be written as $\phi\cos\theta\sin\omega t$ through Q^1 and $\phi\sin\theta\sin\omega t$ through P^1. They are therefore equivalent to a resultant flux $\phi\sin\omega t$ at right angles to a plane making an angle θ with the plane of Q^1 (Fig. 17). If, then, the plane of the search coil S is set at right angles to this plane, *i.e.*, in the direction of the resultant flux, S will have no flux-linkage and no E.M.F. will be produced. Thus zero signals are obtained when the plane of coil S makes the same angle θ with the plane of coil P^1, as the direction of the incident wave makes with the plane of the loop Q, and so the Bellini-Tosi system enables the direction of a transmitter to be determined.

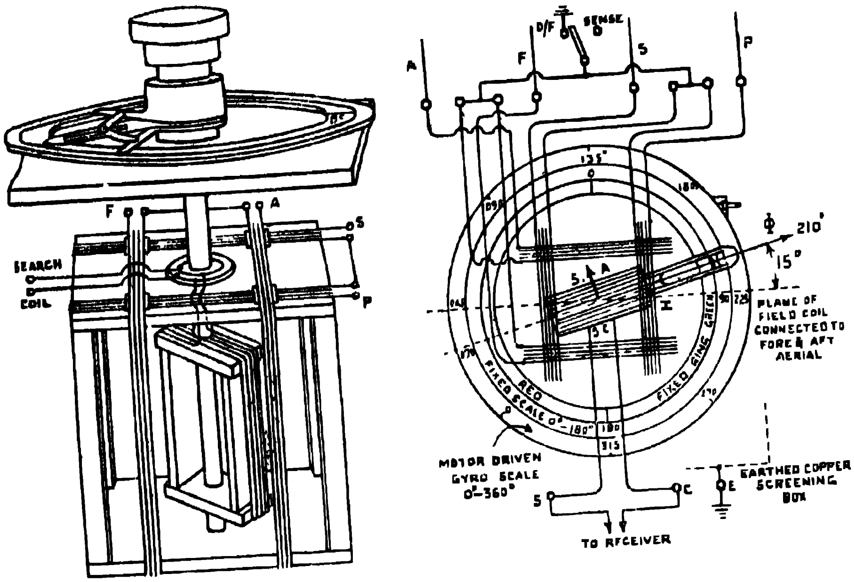
An alternative statement of this result is that the plane of the search coil in the position for zero signals makes the same angle with coil P^1 as the magnetic field of the wave makes with the plane of the loop aerial P to which coil P^1 is connected.

It is interesting to note that the maximum E.M.F. in the search coil S (when coil S is at right angles to the direction of the resultant flux), is independent of the direction of the incident wave, since the maximum flux is always $\phi\sin\omega t$.

14. Radio-goniometers.

In order that the E.M.F. induced in S by mutual inductance with P^1 should be proportional to the cosine of the angle between P^1 and S, as was assumed above, the instantaneous density of flux produced by the current in P^1 must be the same over the whole area of S; and the same applies to Q^1 and S. This is attained by winding the coils P^1 and Q^1 in two halves, as shown in Fig. 18, their distance apart being equal to their mean radius. The instantaneous field due to either coil then has an approximately constant value over the whole region in which the search coil S moves.

Fig. 19 shows a radio-goniometer having a *fixed* search coil (5) and *rotating* field coils (1), (2) and (3) (4). The field coils are placed very accurately at right angles to each other and are rotated in the space enclosed by the search coil. The principle of action is clearly the same as that of the goniometer of Fig. 18. The search coil is formed of two identical windings, the accuracy of the instrument depending on the production of equal magnetic fields by equal currents in the two halves of the coil. Each field coil is split into two halves, the inner ends of each half being connected to earth via a spindle. The outer ends of each field coil are taken to slip rings on the spindle. The



(b)

(a)

FIG. 18.

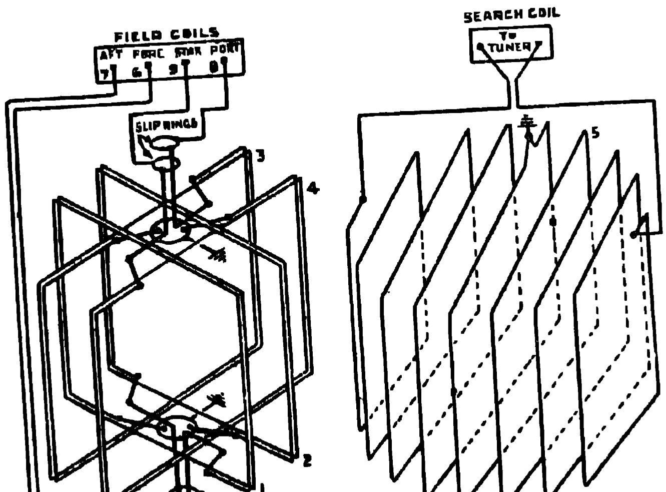


FIG. 19.

field coils are rotated at right angles to their axis inside the search coil by a handle on top of the instrument. Attached to this handle is a pointer which moves over the fixed scale graduated to show relative bearings. Outside the fixed scale is a rotating scale graduated from 0-360°, which is driven by the ship's master gyro compass, enabling true bearings to be read directly from the goniometer.

Untuned aerials are now always used in the Bellini-Tosi system, the desired signal being selected by tuning the search coil, or by any of the methods already described in paragraph 10.

In order to reduce antenna action with Bellini-Tosi aerials, it is usual to join the mid-points of the windings P^1, Q^1 to earth. This is well seen in Fig. 19. When the loops are not unduly large, screened tubes of the type shown for the rotating coil in Fig. 14 are frequently used for the same purpose. Where space is available, it is best to use large loops and a single turn of wire; where space is restricted, loops or crossed frame coils with several turns may be employed, but it must be remembered that increasing the number of turns will not generally compensate for the reduction in size. For multi-turn coils used in a Bellini-Tosi system, four turns are usually found satisfactory, and little improvement is obtained by using a larger number. For best results the inductances of P^1 and Q^1 should be approximately equal to the inductances of the loops or coils.

15. **The Goniometer in a Ship.**—On shore, Bellini-Tosi aerials will normally be set up in the North-South and East-West planes, and the goniometer arranged in any convenient position in the office. A pointer is attached to the moving element, and passes over a scale graduated from 0° to 360°, while the moving part turns through 360°. The pointer is adjusted so that it reads 0° when zero signals are obtained with a wave known to be incident from due North. The true great circle bearing of any transmitter is then given directly by the pointer reading for the position of zero signals. In a ship, the aerials are normally placed with their planes in the fore and aft line, and athwartships. If the goniometer pointer is then adjusted so that it reads 0° in the zero signal position, with a wave known to be incident from right ahead, its readings in direction-finding determinations give bearings relative to the fore and aft line of the ship.

A numerical illustration of this is given in Fig. 18 (*b*), which represents a plan of a Service radio-goniometer. The ship is supposed to be steaming on a course of 135°, and so the scale marking 135°, on the motor-driven gyro scale, is opposite 0° on the fixed goniometer scale. Suppose that the transmitter whose direction is to be found, bears 210° from the ship; the magnetic component of the wave is then in a direction 120°, making an angle of 15° with the fore and aft loop aerial. Hence the resultant flux in the goniometer makes an angle of 15° with the plane of the field coil connected to the fore and aft aerial (Fig. 17).

Zero E.M.F. will then be produced in the search coil if it is placed as shown, its plane making an angle of 15° with the plane of the coil connected to the terminals FA. The D/F pointer, which is permanently fixed in the plane of the search coil (the angle halving device not being in use), then indicates green (or starboard) 75° on the relative bearing scale, and 210° on the gyro compass scale.

The determination of a bearing, after all adjustments have been made, thus consists of the simple operation of rotating the search coil until zero signals are obtained in the telephones, and of reading off the bearing on the gyro compass scale of the D/F pointer. Zero signals would also be obtained if the search coil were rotated through 180°; the pointer would then indicate 030°, the reciprocal of the correct bearing. The reciprocal should be taken as a check if time permits.

For frequencies on which sensefinding is practicable, the determination of "sense," which eliminates the 180° ambiguity, may either precede or follow the determination of a bearing; in the Service, operators are trained to take the bearing first (*cf.* paragraph 35).

16. **The Fixed Crossed Frame Coil Direction Finder.**—Normally Bellini-Tosi aerials as fitted in ships, consist of large loops. If the size of the loops is cut down, the signal strength on frequencies up to 600 kc./s. will suffer proportionally. On higher frequencies, quite small loops give satisfactory results up to about 15,000 kc./s., if attached to suitable circuits and suitably placed. The fixed crossed frame coil direction finder, Fig. 20, is therefore to be regarded as an adaptation of the normal Bellini-Tosi system for use principally on higher frequencies. Its method of action is exactly the same.

Much greater care is, however, required in the design of the component parts. The frames may consist of crossed loops of two turns of wire, each with an area of about 9 sq. ft. The planes of the loops must be set very accurately at right angles to each other. The tuning and pre-selector gear, which has already been described in paragraph 10, may be applied to the search coil. Owing, however, to the reduction in signal strength, a balanced stage of R/F amplification will usually be required before the beat rectfier described in paragraph 10. The electrical balance of the two aerial circuits must be very carefully adjusted.

Such small loops as these are highly inefficient at medium frequenceis, and their use necessitates a specially designed radio-goniometer having a high coefficient of coupling between the search coil and the field coils.

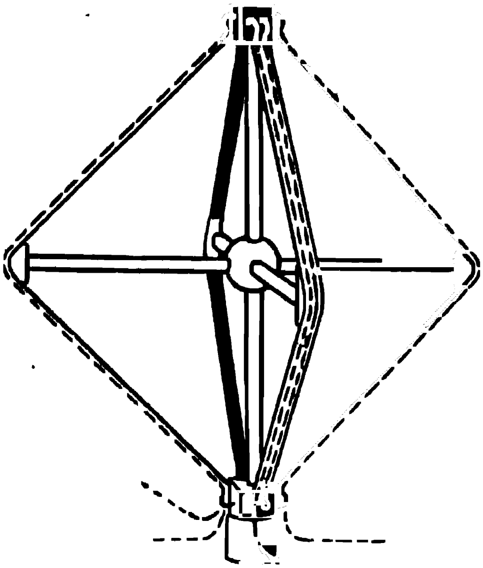


FIG. 20.

17. **Comparison between Rotating Coil and Bellini-Tosi Systems.**—A rotating coil and a Bellini-Tosi system are electrically equivalent. When fitted on shore they give the same result, and there is no question of one system being more accurate than the other. In a ship a difference in their performance is often found, but this can always be shown to be due to the fact that Bellini-Tosi aerials can, in general, be made larger than rotating coils. With aerials giving equivalent signal strengths, in positions that are equally favourable, there is no difference in performance provided the apparatus is well maintained.

The rotating coil, provided it can be fitted in a suitable position, has the merit of simplicity from an electrical point of view. Bad contacts, or other defects, result in a loss of signal strength, and possibly no bearing being obtained, but do not easily give rise to false bearings. Moreover, the coil can be tuned directly by a condenser in many cases, the full resonant voltage thus obtained being available for direct application to the amplifier. There is not the loss in signal strength which is produced in the Bellini-Tosi system by passing the signals across the mutual inductance link between the search coil and the field coils of the goniometer. For this reason it is necessary to make the Bellini-Tosi loops larger in area than the corresponding frame coil. To correspond with a rotating coil of circular section and of diameter 4 ft., it is necessary to have crossed Bellini-Tosi frames of size about 7 ft. square, in order to have signals of equal strength. When large loops offer no disadvantage, and in the case of D/F at higher frequencies using small fixed crossed frames, the Bellini-Tosi system has the merit that the only element to be rotated is a light coil. For rapid working this is an advantage, particularly at H/F, when signals are fading. The disability of the system is the necessity for perfect electrical balance between two separate aerial circuits. This balance is at once destroyed by a bad contact in one circuit and a bearing up to 90° in error can result. Moreover, the balance progressively becomes more difficult to achieve as attempts are made to design direction-finders for work on higher frequencies.

18. **Instrumental Design.**—In all direction-finders it is essential that the whole of the energy received by the amplifier, reaches it through the aerials in a known and definite manner. For this reason, every part of the apparatus, including all tuners, amplifiers, batteries, &c., must be so well screened from the direct action of the wave that it does not pick up an E.M.F. capable of giving an appreciable signal, *i.e.*, not more than ½ per cent. of the signal picked up by the aerials.

Apart from those due to imperfect screening from the direct action of the wave, errors may sometimes be produced by undesirable couplings between instruments inside the office, or by unsuitable circuit arrangements. Instrumental defects frequently have the characteristic that the E.M.F. due to them does not change appreciably as the frame or search coil is rotated. The result of this is that opposite minima are either not exactly 180° apart, or are of different quality, *i.e.*, one is more blurred than the other. Whenever symptoms of this kind are present, suspicion should be directed to :—

(*a*) Pick-up by the receiver directly from the wave or from other aerials coming into the office ;

(*b*) Antenna action of the aerials ; or

(*c*) Defects in the circuits.

From this point of view it may be said that the rotating coil is less liable to instrumental errors than the Bellini-Tosi system.

With both systems it is necessary to ensure that the circuit to which the rotating coil (or search coil) is connected, is symmetrical about the two ends of the coil. The higher the frequency the more essential is perfect symmetry.

With Bellini-Tosi systems the two aerial systems should have identical electric constants, except in so far as they are deliberately altered to allow for the effect of local structures. The goniometer must be completely screened from all other instruments. To make certain that it is itself free from error, it is subjected to a number of tests before being issued for service. These tests ensure that when the winding which is joined to the fore and aft aerial is alone is use, two zeros are obtained at 0° and 180°, and that when the other winding is alone in use the zeros are exactly at 90° and 270°. The accuracy of the instrument at intermediate points is ensured by a proper choice of dimensions for the search coil.

Sometimes signals are so weak, or the bearing so ill defined, that it is not possible to read the position of the "zero" directly on the scale. Recourse may then be had to an angle-dividing device which may be fitted to goniometers and training units. Some practice is needed to attain proficiency in judging the two positions in which the signal is of equal intensity, especially if interference is present ; for the interference will probably not be of the same strength for two positions of the coil equally inclined to the direction of the signal whose bearing is required. The angle divider is intended to reduce computation, and it should be remembered that when a cam corrector is also fitted, the reading of the cam pointer will give a corrected true bearing without further trouble.

19. **Determination of Sense.**— The ambiguity of the bearing obtained by the direction finders described above has already been indicated. For any direction of the incident wave, zero signals are found in two positions 180° apart, using either a rotating coil or the search coil of a fixed loop system, and correspond respectively to the bearing of the transmitter and its reciprocal bearing. It is therefore necessary to devise some means of determining which of the two pointer readings is the correct one. Instruments for this purpose are called "sensefinders." The usual principle on which they operate is as follows ; according as the transmitter lies on a certain bearing or its reciprocal, the loop E.M F., when the plane of the loop is in the direction of propagation, will have the same amplitude, but will be altered in phase by 180°. If at any instant it was directed clockwise round the loop for the one bearing, then for a transmitter on the reciprocal bearing it would be counter-clockwise round the loop at the same instant. This, however, does not apply to the E.M.F. in a vertical aerial, the amplitude and phase of which are unaffected by the direction of the transmitter.

Suppose, therefore, that in addition to the rotating coil we have an open aerial which can be coupled to the receiver. The wave induces an E.M.F. in this aerial at the same moment as an E.M.F. is induced in the loop, but the E M.F. which is induced in the open aerial does not vary with the direction of the wave. The E.M.F. induced in the loop is a maximum when the loop is in the direction of propagation of the wave, and changes sign when the coil is rotated through 180°. Therefore, if these two E.M.Fs., one from the loop and one from the aerial, can be made to act on the receiver at the same time, it will be seen that for one position of the coil, viz., when it is pointing towards the transmitting station, the two E.M.Fs. will reinforce each other, but that the two E.M.Fs. will be opposed when the coil is rotated through 180° from this position. If, now, we make the two E.M.Fs. equal in magnitude, they will cancel each other when opposed, and a zero will be obtained. In the position of the coil 180° from this there will be a maximum due to the combined effect of the two E.M.Fs. assisting each other. When the coil is at right angles to the direction of the wave, the E.M.F. from the open aerial will still be operative, so that, instead of a zero, the signal will continue to be heard. In fact, there is only one position for a zero during a rotation of the coil through 360°. Fig. 21 explains this matter in further detail. Assume the loop to be set to the position of maximum instantaneous E.M.F. and that this is numerically equal to two millivolts. Fig. 21 (*a*) shows this at an instant of time. If another two millivolts is "injected" in the same direction at this instant, the maximum voltage becomes four millivolts. Both the injected and the loop E.M.Fs. undergo their sinusoidal variations, varying from 0-4 millivolts, but always remaining in phase. Fig. 21 (*b*)

shows the loop on the opposite bearing, with the injected and loop E.M.Fs. in phase opposition. As they undergo their sinusoidal variations they will always remain in opposition and so cancel each other. At any angle θ, the loop E.M.F. can still instantaneously be in phase with the injected E.M.F., though numerically it will be smaller, and for $\theta=90°$ the loop E.M.F. is zero. When θ becomes greater than 90° the loop E M.F. again grows in value but is in anti-phase to the injected E.M.F. and the numerical value of the maximum E.M.F. accordingly decreases. Everywhere in the second and third quadrants the loop E.M.F. is instantaneously in anti-phase to the injected E.M.F., and if their magnitudes are equal, a perfect zero will be formed when $\theta=180°$.

In order that the amplitudes of the vertical and loop E.M.Fs. may be added or subtracted in this manner, these E.M.Fs. must either be in phase or 180° out of phase. But it has been pointed out in paragraph 3, that the loop E.M.F. is 90° out of phase with the vertical E.M.F. The actual effects required to be additive or subtractive are not the E.M.Fs. themselves, but the P.D.'s they produce across the first stage of the amplifier. Hence, in order that sensefinding may be possible, **an alteration in phase by 90° of one effect relative to the other must be brought about.** Methods of accomplishing this are explained below. If we assume for the moment that it has

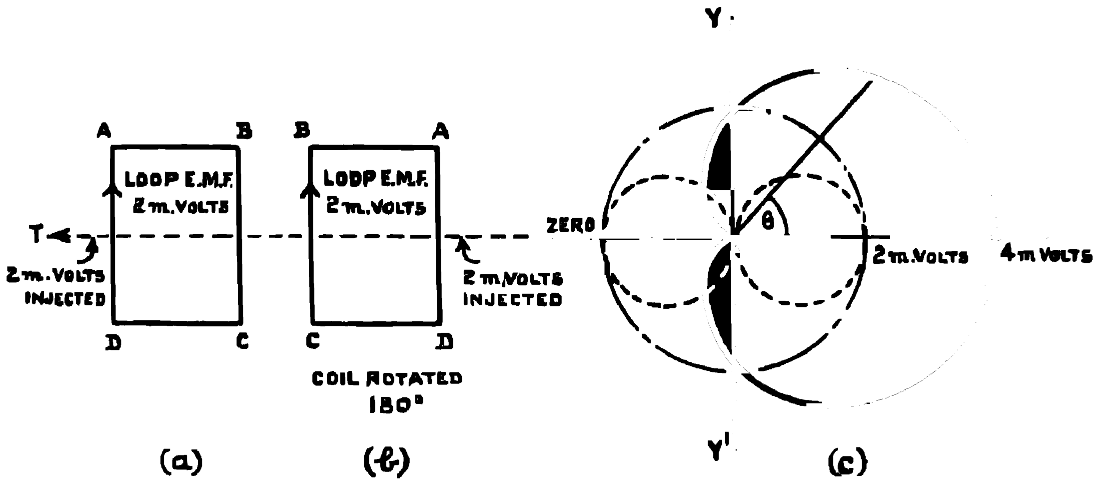


FIG. 21.

been successfully achieved, the variation of resultant input P.D. to the amplifier, due to both E.M.Fs. as the coil is rotated, may be obtained graphically, as in Fig. 21 (*c*).

The P.D. due to the loop E M.F. is represented by the figure of eight diagram (para. 3), and that due to the vertical aerial by a circle. The amplitude of the vertical P.D. is adjusted to equality with the maximum P.D. produced by the loop E.M.F. In other words, the radius of the vertical aerial circle is equal to the diameter of either circle of the loop figure of eight diagram. In the right-hand circle of the loop polar diagram of reception, the P.D. is supposed to be in phase with the vertical P.D., and to be 180° out of phase in the left-hand circle. Radii vectors to the right of YY^1 are therefore added, and the difference is taken of those to the left. Thus the variation of the resultant P.D. with the direction of the loop is represented by the heart-shaped curve, which is called a "cardioid." It will be seen that this curve has only one zero value, and so there is only one orientation of the rotating coil in which zero signals are obtained. It is therefore possible to distinguish between the correct bearing and its reciprocal. The zero will be either in the direction of the transmitter or its reciprocal direction, according to the phases of the two circles of the figure of eight diagram, compared with the phase of the vertical aerial circle. This can be determined for a known bearing, and the pointer set accordingly. It will then give the correct reading for any bearing of a transmitter.

It will be seen that the zero of the cardioid is at 90° from either of the zeros obtained with the

frame coil alone. Provided that the cardioid is perfect in shape, there is no reason why it should not be used directly for taking bearings without ambiguity. But in practice, and particularly in a ship, it is found to be very difficult to obtain a cardioid sufficiently perfect to give accurate bearings in all cases. Another disadvantage is that the change is signal strength as the coil is rotated in the neighbourhood of the cardioid zero, is much less than the corresponding change on the figure of eight diagram. It is therefore usual to take the accurate bearing using the figure of eight characteristic, and to switch in a separate aerial to give a cardioid for the purpose of determining "sense." As the zero of the cardioid is 90° from either zero of the figure of eight, a separate pointer [*e.g.*, the sense arrow S.A. in Fig 18 (*b*)] might be fitted for sense determination. In practice, the operator first obtains the direction with extreme accuracy in the D/F position. He then switches in the sensefinder. It is usual to arrange that if signals DECREASE in strength when the rotating coil or goniometer is turned in a CLOCKWISE direction, the minimum on which the pointer rested indicates the true bearing of the station. If the signals INCREASE, the reciprocal bearing of the station is indicated. Clearly, this is only a convenient rule for rapid operation, and reversal of the electrical connections will reverse the rule. There is, however, a further point that whereas one may fairly easily obtain an increase in signal strength, it is not so easy to obtain a decrease in it unless the sensefinder is working satisfactorily—a matter of considerable importance.

If the E.M.F. at the amplifier due to the vertical aerial is not exactly equal to that due to the rotating coil, the cardioid is distorted.

The resultant polar diagrams of reception for the cases when the vertical aerial E.M.F. is less and greater than the loop E.M.F. are shown in Fig. 22 (*a*) and (*b*) respectively. In Fig. 22 (*a*), two zeros are obtained at equal angles from the correct direction of the transmitter, the size of the angles depending on the ratio of the E.M.Fs. In Fig. 22 (*b*), a minimum of sound is obtained, instead of a

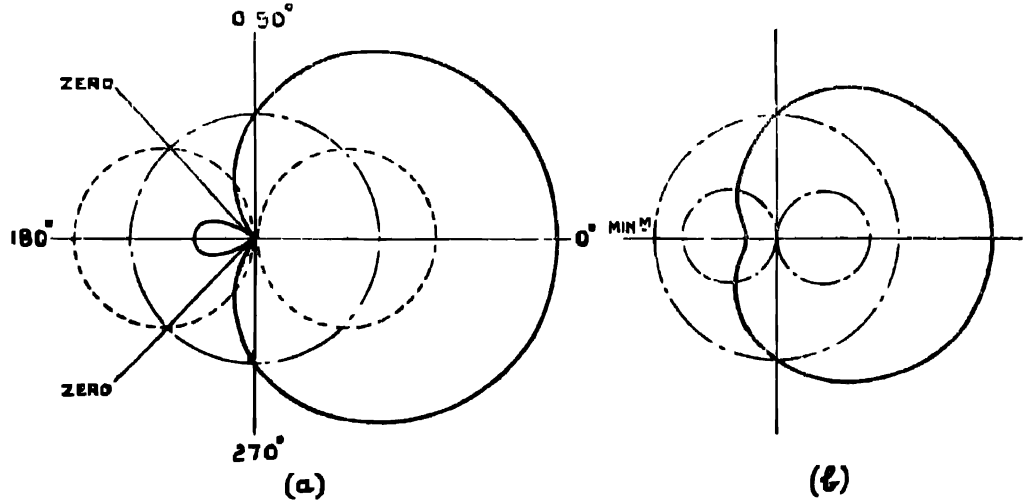


FIG. 22.

zero in the correct position. When the direction of the transmitter has previously been obtained and only its sense is required, these diagrams show that, provided the inequality of loop and vertical E.M.Fs. is small, there is sufficient difference between the maximum signal and the signal on the reciprocal bearing to allow of sense determination.

★20. **Mathematical Note.**—It should be observed that, plotted in polar co-ordinates, the result $E_\theta = E_L \cos\theta$ produces one circle and not two. The figure of eight diagram is, in fact, only a convenient representation of the two quantities, " phase " and " magnitude." The figure of eight diagram is the polar graph of $E_\theta = \sqrt{E_L^2 \cos^2\theta}$.

Further, the polar graphs given by $r_1 = b$ and $r_2 = a\cos\theta$, where a and b are constants, are both circles, the pole being the only difference. The graph is that of Fig. 21 (*c*) without the left-hand circle of the figure of eight diagram. If these two graphs represent the vertical aerial E.M.F. and the loop E.M.F. respectively, then their sum is given by

$$R = a\cos\theta + b.$$

In the general case when a and b are unequal, a " limaçon " results. When a is greater than b, Fig. 22 (a) results, but the small loop along the 180° line should be turned through 180° so as to be along the 0° direction. It follows, therefore, that this figure as drawn is also a convenient representation of the quantities, phase and magnitude, as in the case of Fig. 3.

When $a=b$, the limaçon becomes a cardioid with a perfect zero.

When a is less than b, the limaçon of Fig. 22 (b) results.

21. **Sensefinder Design.**—The problem of sensefinder design therefore reduces to :—

(a) Producing a change in phase of 90° in the effect of the vertical aerial E.M.F. relative to that of the loop E.M.F.

(b) Adjusting the effects to equality in magnitude.

One form of sensefinder is shown in Fig. 23. The vertical aerial is tuned by the inductance L_1. The current in L_1 is therefore in phase with the E.M.F. produced by the wave in the vertical aerial, and so in phase with the wave itself. L_1 is mutually coupled to the inductances L_2 in the loop circuit. The E.M.F. induced into L_2 from L_1 lags or leads by 90° on the current in L_1, and is therefore 90° out of phase with the incident wave. Hence it is either in phase or in anti-phase with the loop E.M.F. produced by the wave, and the two effects can be added. The value of the mutually-induced E.M.F. is made equal to that of the loop E.M.F. by adjusting the mutual coupling.

In Fig. 23 an aerial is shown combined with a rotating frame coil, but the same result may equally well be obtained when using Bellini-Tosi aerials by connecting the search coil of a goniometer in place of the rotating coil.

★22. **Mathematical Analysis.**—The operation of the sensefinder circuit of Fig. 23 may be more directly analysed as follows.

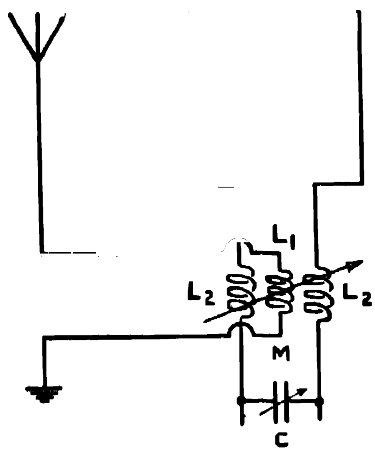


FIG. 23.

Suppose $\mathcal{E}\sin\omega t$ to be the electric field of the wave incident at an angle θ from true North. If the plane of the loop is set at an angle α from true North, the angle between the loop and the direction of propagation is $(\theta-\alpha)$ and the loop E.M.F. is

$$\frac{\omega\mathcal{E}A}{c}\cos(\theta-\alpha)\cos\omega t$$

(paragraph 2).

The E.M.F. in the open aerial is $\mathcal{E}h\sin\omega t$, where h is its effective height. Since the aerial is tuned, the current flowing in it is $\frac{\mathcal{E}h\sin\omega t}{R}$, where R is its total resistance. If the mutual inductance between L_1 and L_2 is M, the E.M.F. induced into L_2 is therefore $\frac{\omega M\mathcal{E}h\cos\omega t}{R}$. The total E.M.F. round the loop is therefore

$$\frac{\omega\mathcal{E}A}{c}\cos(\theta-\alpha)\cos\omega t+\frac{\omega M\mathcal{E}h}{R}\cos\omega t$$

$$=\left[\frac{\omega\mathcal{E}A}{c}\cos(\theta-\alpha)+\frac{\omega M\mathcal{E}}{R}\right]\cos\omega t,$$

and is of the form $(a\cos\theta+b)\cos\omega t$ (paragraph 20).

If M is now adjusted so that $\frac{\omega\mathcal{E}A}{c}=\frac{\omega M\mathcal{E}h}{R}$, the total loop E.M.F. becomes $\frac{\omega\mathcal{E}A}{c}[\cos(\theta-\alpha)+1]\cos\omega t$.

This expression shows how the signal strength varies for different values of α. Between 0° and 360° it has only one zero value, viz., when $\cos(\theta-\alpha)=-1$ or $\alpha=\theta\pm180°$. For the value of α 180° different from this, *i.e.*, $\alpha=\theta$, the signal strength is a maximum, the amplitude of the E.M.F. being $2\omega\mathcal{E}A/c$ (since $\cos(\theta-\alpha)=1$). The horizontal polar diagram obtained by plotting total loop E.M.F. against α is, of course, a cardioid.

23. A Service Sense Finder.—A sensefinder which avoids the necessity of tuning a circuit to resonance, is shown in Fig. 24. In this arrangement use is made of the properties of a valve to give a current in the anode circuit which is in phase with the voltage applied between grid and filament.

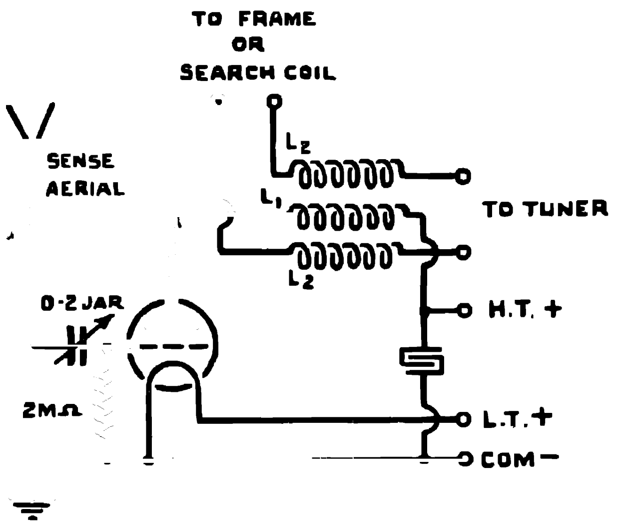


FIG. 24.

The variable condenser of maximum value about 0·2 jar, is chosen so that its reactance will be small compared with the grid-filament resistance, over the range of frequencies for which the sense-finder is employed. This variable condenser may be termed the "sense aerial coupling condenser," and since it also controls the input between grid and filament, it may be regarded as being something like a "volume control." The grid filament resistance is of the order of 2 megohms. Under these conditions the current in the vertical aerial is approximately in phase with the field of the wave. The grid filament input being in phase with the current is therefore approximately in phase with the field of the wave. The anode current is practically in phase with the grid voltage and the E.M.F. induced into L_2 will be 90° out of phase with the current flowing in L_1, and accordingly it will be 90° out of phase with the field of the wave. It is therefore in phase or in anti-phase with the loop E.M.F , and a cardioid diagram of reception can be obtained by suitably adjusting the variable condenser. The circuit has the advantage that the correct phase relation is maintained over a wide range of frequencies without the trouble of adjusting a tuned circuit.

Sense finding becomes progressively more difficult as the frequency is raised, and, in general, it may be said to be satisfactory only up to about 600 kc./s.

Sometimes the separate aerial can be dispensed with by making use of the antenna action of the loops themselves. This is most easily arranged with the large aerials used in Bellini-Tosi systems. In the "sense" position, the electrical mid-points of the field coils of the goniometer are joined to earth through a circuit coupled to the receiver, so as to give a voltage in the correct phase. It can then be arranged that the loops act at the same time in two distinct ways, first as loops with circulating currents flowing round them, and secondly as vertical aerials giving antenna currents down the lead to earth. Sense finders working on this principle are much more complicated in action and design than the simple sense finder described above, and no account of these more complicated instruments will be given here.

24. Errors in Direction Finding.—The errors common to all systems of D/F may be broadly classified into :—

(*a*) Errors which have been accumulated during the passage of the wave.
(*b*) Errors due to the site of the direction finder.
(*c*) Instrumental errors.

It is proposed to consider the first two classes in some detail, using the following subdivisional headings :—

(*a*) *Errors due to the wave.*—These include : (i) night effect ; (ii) convergency correction ; and (iii) shore effect.
(*b*) *Errors due to the gear being fitted in a ship.*—These include : (i) quadrantal error ; and (ii) semi-circular error.

The effect and causes of these errors will now be considered in the above order.

25. **Night Effect.**—The theory of determining the direction of a transmitter by the methods described above, has been based on the assumption that the wave is propagated in a direction parallel to the earth's surface, with its electric field vertical and its magnetic field horizontal. This is approximately the case for the direct or ground radiation from a transmitter, but is not so where reception is either partly or wholly due to the indirect radiation, which has travelled up to the ionised parts of the atmosphere and has been bent round so that it returns to earth at some distance from the transmitter. The effect of such radiation on direction finding apparatus will now be considered.

Suppose that the plane of the loop is set at right angles to the vertical plane in which the wave is supposed to be travelling. If the wave were propagated horizontally, no resultant E.M.F. would act round the loop in this position, and zero signals would be obtained. If, however, the direction of the wave is inclined to the horizontal, cases may arise in which a signal will be heard. One such case may be due to the fact that the plane of polarisation of the wave has been rotated in its passage through the ionosphere. The electric field is then no longer in the vertical plane of travel of the wave, that is to say, the vertical plane through transmitter and receiver. This is shown in Fig. 25 where XOZ is the plane of travel, and $\mathcal{X}'$ is shown making an angle with that plane. It is obvious that $\mathcal{X}'$ can be resolved into two components OX in the plane of travel, and at right angles to the direction of propagation, and OY horizontal and at right angles to OX in the plane perpendicular to the direction of propagation. Suppose, now, that the wave is incident at an angle to the horizon, as in Fig. 26.

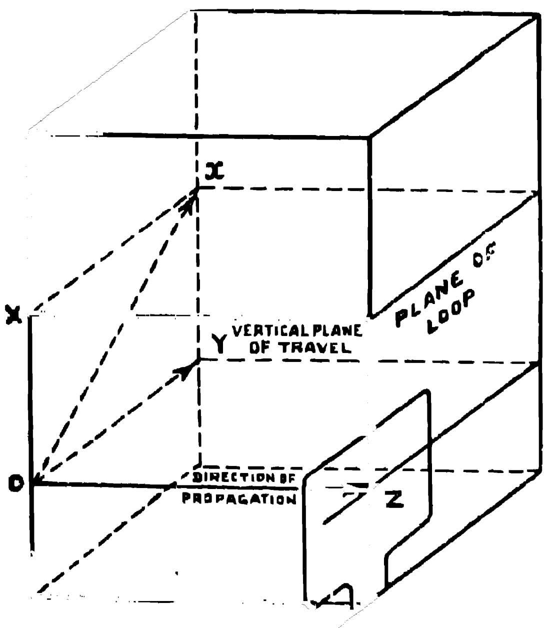


FIG. 25

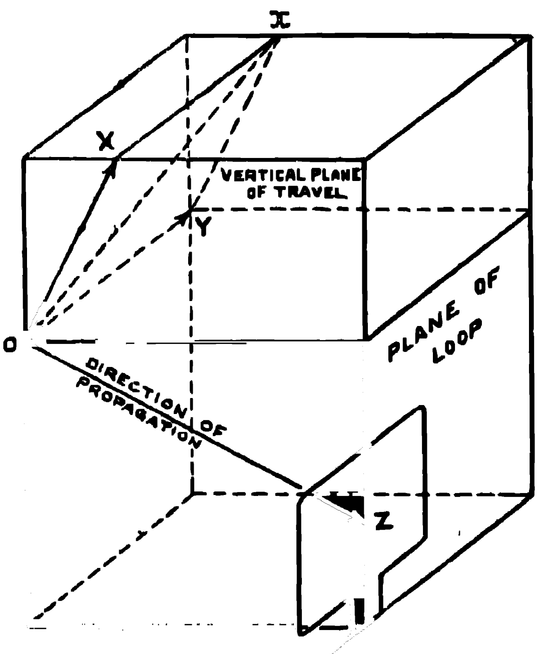


FIG. 26.

The direction of OY is unaltered, and OX, although no longer vertical, is still in the vertical plane of travel. OX therefore produces equal E.M.Fs. in the vertical sides of the loop and no E.M.Fs. along the horizontal sides, and so gives no resultant loop E.M.F. OY produces no E.M.Fs. along the vertical sides, since it is horizontal. It will, however, produce E.M.Fs. along the horizontal sides, and, since it reaches the top of the loop before the bottom, these E.M.Fs. though equal in amplitude if the loop is symmetrical, differ in phase, and so give rise to a resultant E.M.F. round the loop. The loop, in fact, behaves towards the field OY as a loop pointing to the transmitter would

behave to OX when the wave is travelling horizontally. Thus a signal is heard in the position of the loop where a zero should be obtained, and an error is introduced into the determination of direction.

It should be noted that the inclined direction of travel of the wave does not in itself give rise to this particular error, provided the electric field remains in the plane of travel. The error arises when this field has a horizontal component not in the plane of travel. More complex changes in the nature of the wave, such as circular polarisation, will also produce errors in direction determinations. For simplicity, the ray reflected from the earth's surface at the loop has been neglected in the above discussion, but, obviously, its contribution to the loop E.M.F. should also be taken into account.

Near a transmitting station the indirect ray is weak or non-existent on all frequencies, and good bearings can then be obtained. On low and medium frequencies the indirect ray is relatively weak during daylight hours for transmissions over sea up to 100 miles and sometimes further. Good bearings are then possible, with an accuracy of $\pm 2°$. At night the indirect ray becomes important, though relatively less important over sea than over land, because the direct ray is less attenuated over sea, and therefore forms a larger percentage of the whole signal received. At night the reliable range is reduced to about 25 miles.

It is the ratio of the intensity of the indirect to the direct ray in the total received signal, which determines the liability to error of loop direction finders. Since this is greater at night for the usual D/F frequencies and distances, such errors are then most common and of largest amount, whence the name "night effect" given to this phenomenon. It is often found that this effect is at its worst within an hour either side of sunrise and sunset, when the changes in the state of ionisation of the upper atmosphere are particularly violent. Reliable bearings can seldom be taken at these times. As it is impossible for any correction to be applied for this error, it is essential that the risk of its presence should be understood. It is usually shown by an abnormal value of the semicircular corrector (para. 34). The value of the error may be as much as 30°.

Where communication is maintained by using high frequencies, the indirect ray will probably predominate throughout the whole day. The possibility of "night effect" always exists, and the term is a misnomer. It has been shown that night effect errors are due to the voltages induced in the horizontal parts of the loop. The Adcock direction finder, to be described later, aims at reducing these errors by removing the top horizontal sides of the loop, and by screening the lower horizontal leads to the receiver, so that no E.M.F. can possibly be induced in either by an incident wave, no matter what may be its state of polarisation. Such a direction finder could therefore be used on high frequencies.

"Night effect" is also observed when signals are received from an aircraft radiating with a trailing aerial. The waves arrive on the D/F loop in a polarised condition and in a downcoming direction, and give rise to the effect usually known as **aeroplane error.**

26. **Aeroplane Error.**—When an aircraft is transmitting on a trailing aerial, an error arises which is serious at short ranges. The error is due to the fact that the bearing of the machine, taken by a frame aerial, actually indicates the direction of the point where the line of trail of the aerial, if projected, would meet the ground.

Thus, the error is dependent on :—

(*a*) The height of the aircraft.
(*b*) Its direction of flight.
(*c*) Distance.
(*d*) Trailing angle of aerial.

In connection with (*b*) the *error* will be *greatest* when the line of flight is at right angles to the great circle connecting the D/F station and the aircraft, and zero when the direction of the flight is on this great circle, either to or from the station. It is also interesting to note that if the altitude of the aircraft is such that the production of the line of the trailing aerial meets the ground at the

position of the receiver, no signals will be heard : this can only occur when the aeroplane is proceeding directly away from the ship and is instantaneously at the requisite position.

As an example, assuming an effective trail angle of 30°, an altitude of 10,000 ft., and a course at right angles to the line joining aircraft and receiver, the error in position would be $3\frac{1}{4}$ miles behind the aeroplane, and, at a distance of 10 miles, the error in bearing would be 20° (approx.)—*See* Fig. 27.

The error is due to the same cause as night effect and may be explained as follows with reference to Fig. 26.

The vertically polarised field will produce no resultant E.M.F. in the vertical sides of the loop. The horizontally polarised field will, by virtue of its angle of descent, produce a component in the horizontal sides of the loop.

In order, therefore, to obtain zero signals, the loop must be orientated so that there is the same phase difference between the vertical sides, due to direction, as there is in the horizontal sides due to the downcoming horizontally polarised component of the radiation. The two E.M.Fs., if the loop is correctly turned, will cancel out, but the direction indicated will be in error.

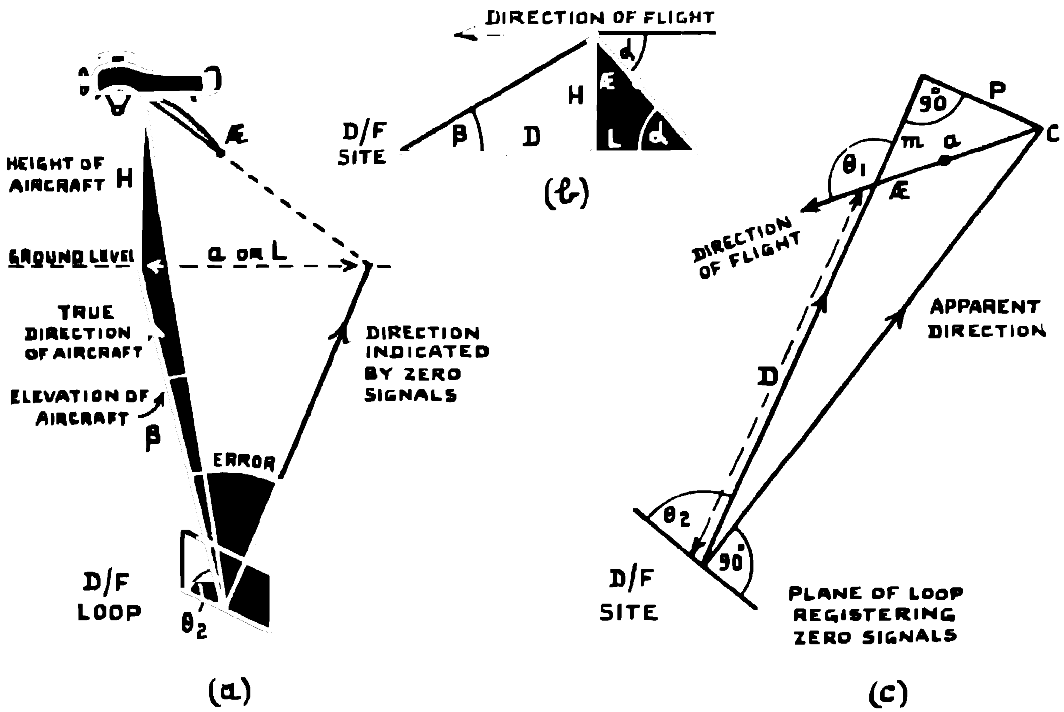


FIG. 27.

Given constant height and speed for any particular aerial, the error can be expressed as distance behind the aircraft per 1,000 ft. of height, an amount which always can be allowed for, irrespective of the course flown.

Fig. 27 (*a*) shows the error which the orientation of the D/F loop produces.

Fig. 27 (*b*) is a projection in the vertical plane, β is the angle of elevation of the aircraft, and is the angle of trail of the aerial.

Fig. 27 (*c*) is a projection on the horizontal plane.

θ_1 is the course flown with respect to the great circle containing the aeroplane and the loop.

θ_2 is the orientation of the loop with respect to the same great circle.

The effect of the error will be of importance when a D/F bearing is used for "homing" an

aircraft. In this case successive alterations of course to the reciprocal of successive D/F bearings will bring the aircraft to a bearing on which the error is practically zero. D/F bearings are signalled as " sense determined " or " not sense determined," and in the case of the latter, aircraft must turn and fly at right angles to the signalled bearing for long enough to ascertain which way the bearing is " drawing " This will resolve any ambiguity of " sense," or detect any error in the signalled " sense."

The errors due to wind, drift of aerial, irregularities of terrain and earth curvature have not been considered, but there is reason to believe that their effect is small.

★ 27 **Mathematical Analysis.**—By equating the E.M.Fs developed in the horizontal and vertical sides of the loop, it has been found that, when zero signals are reached,

$$\sin\alpha\cos\beta\cos\theta_2 - \cos\alpha\sin\beta\cos(\theta_1-\theta_2) = 0$$

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$$\therefore \quad \frac{\sin\alpha\cos\beta\cos\theta_2}{\cos\alpha\sin\beta} - \frac{\cos\alpha\sin\beta\cos(\theta_1-\theta_2)}{\cos\alpha\sin\beta} = 0$$

$$\therefore \quad \frac{\tan\alpha\cos\theta_2}{\tan\beta} - (\cos\theta_1\cos\theta_2 + \sin\theta_1\sin\theta_2) = 0$$

$$\therefore \quad \cos\theta_2\left(\frac{\tan\alpha}{\tan\beta}\right) - \cos\theta_2\cos\theta_1 = \sin\theta_1\sin\theta_2$$

$$\therefore \quad \frac{\cos\theta_2}{\sin\theta_2}\left(\frac{\tan\alpha}{\tan\beta} - \cos\theta_1\right) = \sin\theta_1$$

$$\therefore \quad \frac{1}{\tan\theta_2} = \frac{\sin\theta_1}{\frac{\tan\alpha}{\tan\beta} - \cos\theta_1}$$

$$\therefore \quad \tan\theta_2 = \frac{\tan\alpha}{\tan\beta\sin\theta_1} - \frac{1}{\tan\theta_1} \qquad \ldots\ldots \quad (2)$$

From Fig. 27 (c)

$$\tan\theta_2 = \frac{D+m}{P} = \frac{D}{P} + \frac{m}{P} = \frac{D}{P} - \frac{1}{\tan\theta_1}$$

$$= \frac{\frac{H}{\tan\beta}}{a\sin\theta_1} - \frac{1}{\tan\theta_1} \qquad \ldots \quad (3)$$

From Fig 27 (b) and (c)

$$\frac{\tan\alpha}{\tan\beta\sin\theta_1} - \frac{1}{\tan\theta_1} = \frac{H}{a\sin\theta_1\tan\beta} - \frac{1}{\tan\theta_1}$$

$$\therefore \quad \frac{H}{a} = \tan\alpha = \frac{H}{L}$$

$$\therefore \quad a = L$$

Therefore, the apparent position of the aircraft is that of the point at which the line of trail of the aerial will meet the ground.

28. **Convergency Correction.**—The direction finder measures the great circle bearing of the received signals. If it is desired to plot the bearing of the transmitter on a Mercator's chart, it will be necessary to apply a correction allowing for the difference between the great circle and mercatorial bearings. This correction is capable of mathematical computation from the formula

$$\tfrac{1}{2}\ \text{d long} \times \sin \text{Mid. lat.}$$

This correction is equal to half the convergency of the two meridians of longitude passing through the transmitter and the receiver. " Convergency " is defined as being the difference between the angles which the great circle passing through the transmitter and the receiver makes with their respective meridians.

The correction is given in tabular form in various publications, or it may be quickly determined with the help of a nomogram (a calculating graph). It is therefore clear that this is to be regarded not as a deviation or D/F error, but as a correction which needs to be applied for certain purposes. In general, for short distances from the transmitting station the correction is small.

This subject is more fully explained in text books on Navigation.

29. **Coastal Deviation, or Shore Effect.**—Wireless waves, being electromagnetic in character, are bent or refracted from their normal path when they pass from one medium to another. A small difference in the conductivity of the surface over which the wave is propagated, a change in the value of the dielectric constant of the medium, or even a difference in the state of the tide, is sufficient to constitute a change of medium, and to give rise to refraction effects. The velocity of a wireless wave over sea water may be up to 5 per cent. greater than its velocity over land. If a wave crosses the coast line at an angle of 20° or less, its direction of travel is appreciably

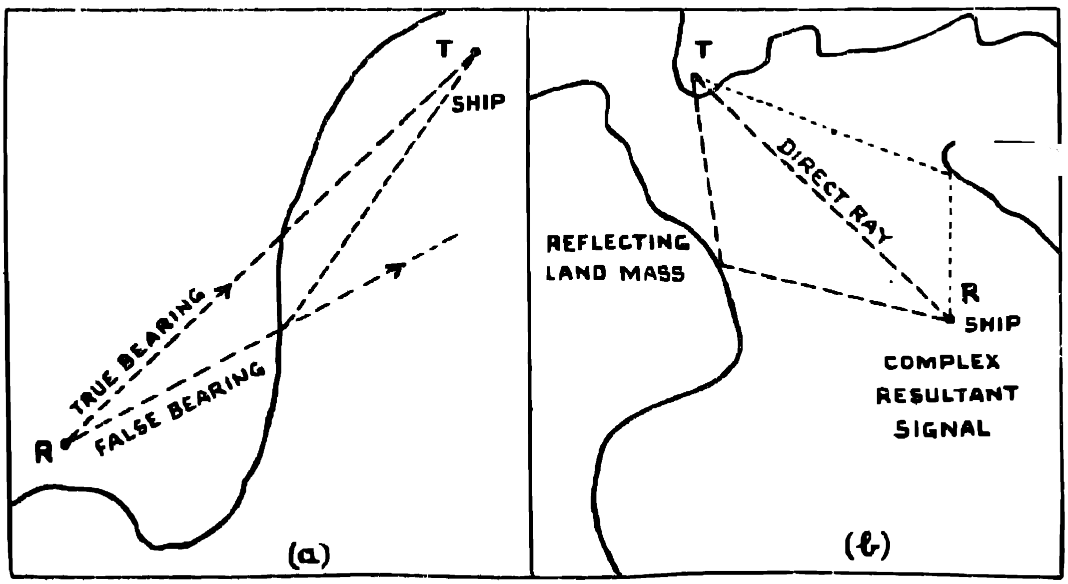


FIG. 28.

altered and the bearing of the wave is no longer the bearing of the transmitter (Fig. 28 (*a*)). The effect is particularly pronounced when either station is near the coast, especially if high land intervenes. As in the case of light rays, the direction of bending is always towards the normal, in the case of waves travelling from one medium to another in which the velocity of propagation is smaller. Reference should always be made to a chart to ascertain whether this cause of error is likely to be present. By analogy with the dispersion of light, the effect would be expected to vary with the frequency of the wave. This is found to be the case for frequencies above 150 kc./s., but below this frequency the refraction appears to be almost independent of the frequency.

In addition to D/F errors produced by refraction effects, incorrect bearings are also frequently obtained under conditions that suggest that reflection may sometimes be a contributory cause

Quite large and otherwise unaccountable errors sometimes are produced when there is land in the neighbourhood of a ship, from which other rays from a transmitter could be reflected (Fig. 28 (*b*)).

Reflection of E.M. waves almost always is accompanied by an alteration in the plane of polarisation, and the resultant signal at the ship is a complex mixture of direct and reflected rays differing in phase and plane of polarisation. The result is a false bearing, which may sometimes be as great as 7° in error. The error is usually big when the base line TR subtends a large angle at the reflecting land mass. It is an error which depends upon the position of the ship, the frequency in use, and the relative bearing of the transmitting station. Unreliable bearings are also sometimes found when there is high land at the back of a shore transmitting station.

It is important to be aware of the possibility of errors of this kind when undertaking the calibration of ship D/F outfits.

30. **Errors Due to the Ship.**—The performance of a direction finder in a ship is found to depend very much upon the position in which it is erected. A direction finder which gives good results when placed in a clear position aloft may, on transference to the upper deck, be found to give very poorly defined zeros and large errors. Moreover, a great deal depends on the frequency. A position for a direction finder which may be quite satisfactory on medium frequencies, is found to be unsatisfactory for high frequencies. On the other hand, if a position can be found which is satisfactory for high frequencies, then an excellent performance is obtained on medium ones.

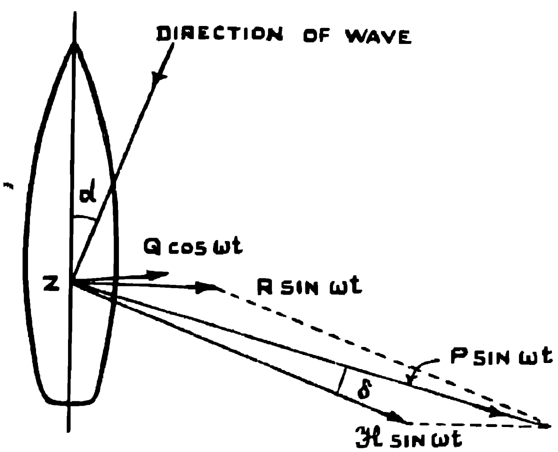


FIG. 29.

In order to examine the matter in greater detail, consider a plane wave which strikes a ship from a direction making an angle α with the fore and aft line (Fig. 29). Alternating currents are induced in the hull of the ship, in the metal work of the superstructure, and in all aerials and other conductors. At a point Z, where we may imagine a direction finder to be situated, each of these currents produces a subsidiary inductive magnetic field which links with the direction finder and may produce error in its determinations. These fields will, in general, be different in magnitude, direction and phase, but they can in any case be resolved into two fields, one in time phase with the field of the wave, and the other in quadrature—90° out of time phase—with the field of the wave. If the alternating magnetic field of the wave is represented by $\mathscr{H} \sin \omega t$, the magnetic forces acting at the point Z can therefore be reduced to—

(i) a component $\mathscr{H} \sin \omega t$ due directly to the wave and directed at right angles to the direction of propagation;

(ii) a component $R \sin \omega t$; and

(iii) a component $Q \cos \omega t$.

In the most general case, R and Q may be at any angle to the fore and aft line, but in practice, especially on medium frequencies, R and Q **are found to be approximately athwartships.** The two fields R and H can be combined to form a resultant P at an angle δ from the direction of $\mathscr{H}$. The angle δ is the deviation due to the metal work of the ship. The effect of the quadrature force Q is to blur the readings more or less (para. 32). In a good position for a direction finder both Q and R will be a small percentage of H. In other words, the field due to the ship will be small compared with the field due directly to the wave. If this is not the case it is difficult to correct for the effect of the ship and good direction finding is impossible.

31. **Ship Field in Phase with the Wave Field.**—This is the field R sin ωt. To a first approximation, this effect of the ship on a direction finder may be considered to be due to the action of a large fore and aft loop, the plane of which is vertical and includes the fore and aft line of the ship.

Such a loop has no current induced in it by a wave incident athwartships and is then without influence on the direction finder. For a wave incident from right ahead, a maximum current is induced in the loop. The subsidiary field due to this current is athwartships and therefore does not link with the direction finder when the latter is on the minimum, so that, once again, the loop produces no apparent deviation of the wave. When, however, the wave is incident on either quarter or bow, the loop has a current induced in it that produces an athwartships field which combines with the direct field of the wave, so that the measured bearing differs from the true bearing by the angle of deviation "δ." A "zero" will be obtained when the plane of the loop is set in the plane of P, and the signal will appear to be **arriving from a direction nearer to the fore and aft direction than is actually the case.** The deviation is a maximum for $\alpha = 45°$, 135°, 225° and 315°, changing sign every 90°—Fig. 30. A deviation of this type is termed "**Quadrantal**" and methods of correcting for it have been employed since the earliest days of direction finding in ships.

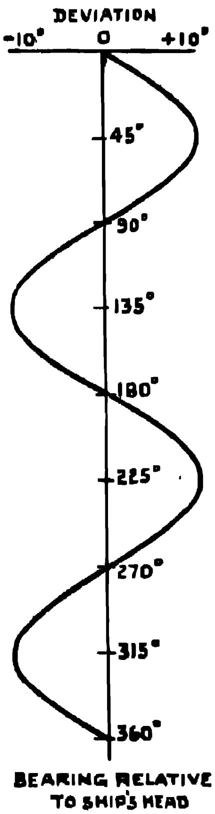


FIG. 30.

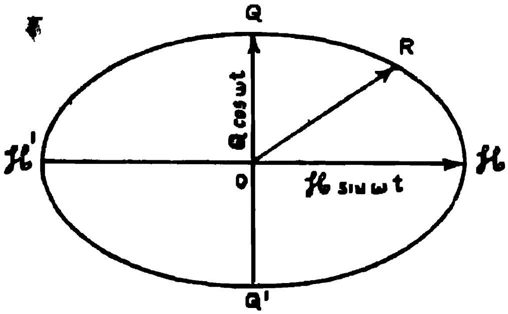
This simple case is found in practice to hold fairly well for medium frequencies when direction finders are fitted on the centre line of a symmetrically built ship, but on higher frequencies and for direction finders unfavourably situated or far from the centre line, it is found that the deviation does not follow this simple type and can only be determined by actual trial. Nevertheless, when the curve of deviation has once been determined, it is possible to apply the requisite correction to an observed D/F bearing and obtain the true direction of arrival of the wave. Mechanical devices for applying this correction automatically have been invented and usually consist of specially made cams. These cams can only be constructed after calibration and usually only hold good over a limited frequency band. They are often used in conjunction with electrical correctors, the latter eliminating the main quadrantal component while the cam corrector takes care of the residual and irregular error.

32. **Ship Field 90° Out of Phase with the Wave Field.**—This is the field Q cos ωt. To a first approximation it can be considered to be due to the vertical conductors in a ship acting as open aerials. All of the vertical metal of a ship may be considered to be replaced by a number of similar vertical aerials spaced along the fore and aft line, their length being small in comparison with $\lambda/4$ and the spacing being some small fraction of λ.

Since all of the aerials in this "array" are being operated below their resonant frequency, the aerial currents and hence the re-radiated inductive magnetic fields will all be approximately 90° out of phase with the field of the wave. Moreover, such an array of miniature transmitting aerials, all operating in phase, will be markedly directional, the re-radiated field being relatively most intense in an athwartships direction, at right angles to the line of the array. (*Cf.* Section on Aerial Arrays.) Since, in general, this field differs in direction and phase from the field of the wave, the two cannot be combined to give a resultant alternating field in any particular direction. The resultant fields turns through every direction from 0° to 360° in one cycle of the components, and so is known as a rotating field.

This effect is most simply understood when the fields are supposed to be at right angles to each other in space, as shown in Fig. 31. Suppose that time is reckoned from the instant

when $\mathcal{H}$ sin ωt has its maximum value $\mathcal{H}$. The value of Q cos ωt is then zero, and so the resultant field is O$\mathcal{H}$. A quarter of a period later, $\mathcal{H}$ sin ωt is zero and Q cos ωt has its maximum value Q. The resultant field is therefore of value OQ in the direction of Q cos ωt. After another quarter of a period, Q cos ωt is again zero, and $\mathcal{H}$ sin ωt has its maximum value in the opposite direction, giving the resultant O$\mathcal{H}'$. Three-quarters of a period from the initial moment, $\mathcal{H}$ sin ωt is zero and Q cos ωt has a maximum negative value. Hence, the resultant is OQ′. At the end of one period the resultant returns to its initial value. At intermediate times the resultant field is the sum of the instantaneous values of the two components, neither of which is zero, and so is in a direction intermediate between the directions of the components. It is shown for an instant in the first quarter period by OR. The ends R of all the instantaneous resultant fields during one cycle form the curve shown in the figure, which is called an ellipse. The resultant field represented by the vector OR sweeps out the area of this ellipse during one cycle of values of the components, and is called an elliptical rotating field. When $\mathcal{H}$ sin ωt and Q cos ωt are not at right angles, their resultant is still an elliptical rotating field, and this is also the case if the effect of P sin ωt is taken into account.



ELLIPTICAL ROTATING FIELD

FIG. 31.

It is obvious that when such a field is present, there is no position of a rotating loop in which a zero signal can be obtained. The effect of the field Q cos ωt is to blur the zeros and render it difficult to obtain precise readings.

It is always desirable so to situate a direction finder that the ratio $Q/\mathcal{H}$ is small. Where this cannot be done, some method of neutralising the effect of Q is advantageous. An open aerial is usually employed. If this is far from resonance, the current induced in it is in phase with Q cos ωt and can be used to neutralise the blurring—para. 34. An open aerial used for this purpose should itself be as far as possible from the metal work of the ship ; otherwise the aerial is itself appreciably influenced by re-radiation and it is found impossible to keep the current in it sufficiently closely in phase with Q cos ωt. It is usual to employ the same aerial for this purpose as that used for sense-finding, care being taken that the aerial circuit is far from resonance when used for "zero clearing."

For the reason given above, the force Q, in the simplest case, is approximately athwartships for all directions of incidence of the wave. Its effect is therefore greatest when it is at right angles to the field due directly to the wave, that is, when the wave is incident on either beam. In other words, blurring is most noticeable for stations on either beam, stations ahead and astern frequently giving very clearly defined readings. It is usual to apply zero clearing by means of a variable magnetic coupling, and it is found that the amount required is approximately zero for waves incident ahead or astern, and changes sign in passing from starboard to port. In other words, it is of a "**semi-circular** character," changing sign every 180°. For this reason, zero clearing devices are sometimes referred to as "semi-circular correctors." The amount of semi-circular correction required is dependent on the precise rig of aerials and superstructures in the ship and, of course, varies for different angles of incidence of the waves. It is, however, determined during calibration and any departure from the amount found necessary during calibration is an indication that conditions have changed, and that to a greater or lesser degree the calibration is invalid. Such a state of affairs may occur at night when there is an ionospheric as well as a ground ray. If it is found that the necessary amount of semi-circular correction is abnormal for the particular bearing in question, it is proper to conclude that night effect is present. A zero clearing device, therefore, has the additional advantage that it provides an indication to the operator that night effect is present. thought it can do nothing to reduce it.

33. **Quadrantal Correctors.**—In order to obtain true bearings without the necessity for applying corrections for quadrantal deviation, a number of quadrantal correctors have been devised.

(1) **Bellini-Tosi Systems.**

(*a*) One loop is set up fore and aft and the other loop athwartships. To counteract the effect of the ship, which tends to make the wave appear to be coming from a direction more nearly right ahead than it is, the fore and aft aerial is intentionally made smaller so that with a transmitting station bearing 45°, the currents flowing in the two primary windings of the goniometer are approximately equal. The goniometer then gives the true bearing 45°, over a wide frequency band.

(*b*) Equality of currents for a wave incident at 45° may also be obtained by adding impedance to the circuit of the fore and aft loop. This takes the form of two equal inductances, one in each leg of the aerial, but inside the office. They are adjusted by trial until the zero is at 45°.

(*c*) A similar result may be obtained by shunting the winding of the goniometer connected to the fore and aft aerial by an inductance. The larger the inductance the less the shunting effect, and it is possible to calculate the amount of inductance required to correct any given amount of deviation. In Service practice it is usual to provide a shunt for both the goniometer winding joined to the fore and aft aerial and that joined to the athwartship aerial, as in Fig. 32; use is made of the requisite shunt. The principal advantage of this method of correcting quadrantal error is that a defective contact in the shunt inductance can at the most give an error equal to the deviation, whereas a defective contact in a series inductance of method (*b*) can lead to an error up to 90°.

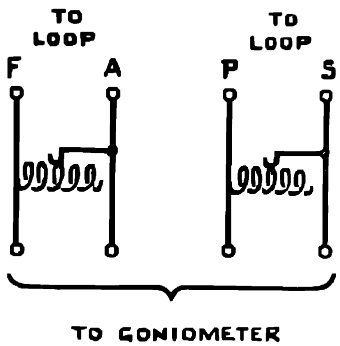


FIG. 32.

(*d*) A cam corrector may be attached to the search coil and used as described below; it may be used in conjunction with (*c*). The reading is usually given by a "floating" pointer, operating independently.

(2) **Rotating Coils.**—The most usual practice is some form of cam corrector by which the pointer is made to lag or lead on the coil so that the readings given by the pointer are true bearings, while the coil itself merely determines the apparent bearing. The advantage of a cam is that it can be so shaped as to correct any type of deviation due to the ship, whether it is of the simple quadrantal or a more complicated type.

The quadrantal deviation due to a ship is usually found to be constant for frequencies corresponding to wave lengths longer than about five times the length of the ship, but increases appreciably on higher frequencies. For this reason it is necessary to adjust quadrantal correctors for the frequency in use at the time, and to find out by trial how the correct adjustment varies with the frequency.

In Bellini-Tosi systems having a long length of cable between the aerials and the office, the behaviour of quadrantal correctors is appreciably modified by the capacity of the cable. They will only behave in a normal manner if the circuit, consisting of the aerial inductance and the cable capacity, is very far from resonance.

34. **Semi-Circular Correctors.**—To compensate for semi-circular effect it is necessary to introduce into the loop an E.M.F. equal and opposite to the ship field which blurs the zeros. From the previous analysis of this error it will be seen that the correcting E.M.F. must be 90° out of phase

with the E.M.F. induced directly in the loop by the incoming wave, *i.e.*, in phase or in anti-phase with the field that causes blurring.

The semi-circular corrector may consist either of a vertical aerial or a loop aerial circuit.

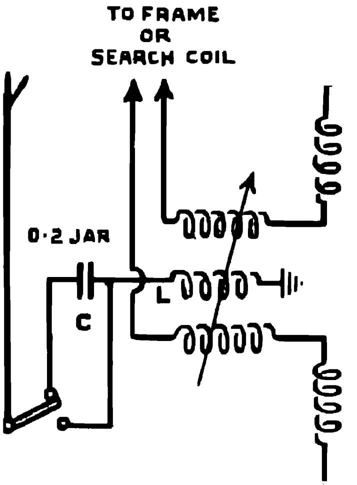


FIG. 33.

A vertical aerial corrector is shown in Fig. 33. The aerial is always operated in a condition which is far from resonance, and so the current flowing in the inductance L is 90° out of phase with the vertical aerial E.M.F., *i.e.*, 90° out of phase with the field of the wave. The mutually induced E.M.F. into the loop circuit is therefore in phase or anti-phase with the field of the wave. The loop E.M.F. directly induced by the wave is 90° out of phase with the wave field. Hence the correcting E.M.F. is 90° out of phase with the directly induced loop E.M.F. It is also 90° out of phase with that produced by a sense-finder.

The amplitude of the correcting E.M.F. is varied by adjusting the mutual coupling. The amount of coupling required varies with the direction of the wave, and *usually*, if it is positive for starboard bearings, it is negative for port bearings, being zero for bearings exactly ahead and astern. Bearings should always be taken first with the coupling of the semi-circular corrector at zero. If a blurred zero is obtained, the corrector may be brought into operation by increasing the coupling. A value of coupling in one direction or the other should quickly be found where the blurring is reduced, and it should be only a matter of careful adjustment of coupling, and the setting of the search coil or frame coil, in order to obtain a perfect zero. When the minimum has been found by use of the corrector, it will be observed that the reciprocal minimum has been blurred still more. It must not be inferred, however, that the perfect zero gives the true direction (*i.e.*, the sense) of the station. With a loop aerial corrector this can sometimes be arranged.

The series condenser is provided for use with a large aerial which may—

(*a*) give too strong signals, or

(*b*) come into resonance within the range of frequencies in use.

In the case of (*a*), the settings of the coupling in the corrector may be too critical if the series condenser is not employed, and in the case of (*b*) the current flowing in the aerial will not be in the right phase, but can be put right by bringing in the series condenser. It is often convenient to use the same vertical aerial for sense finding and for semi-circular correction, since, in general, the two operations are not done simultaneously.

Loop aerial semi-circular corrector circuits are more complicated in action and less commonly used. They will not be described here.

35. **Typical Receivers.**—Fig. 34 shows a simplified diagram of the first stages of a circuit suitable for use at low and medium frequencies, incorporating sense finder, semi-circular corrector, and other features which have been discussed above. The diagram only shows the input of the first balanced stage of radio frequency amplification. In practice this is followed by additional stages, detection, and note magnification.

Fig. 35 shows a simplified diagram of a circuit suitable for use at high frequencies employing beat rectification and having only one further stage of intermediate frequency amplification before detection. The note magnifier is not shown, and no sense finder is incorporated. It is to be understood that in the complete receiver several further stages of intermediate frequency amplification will be required.

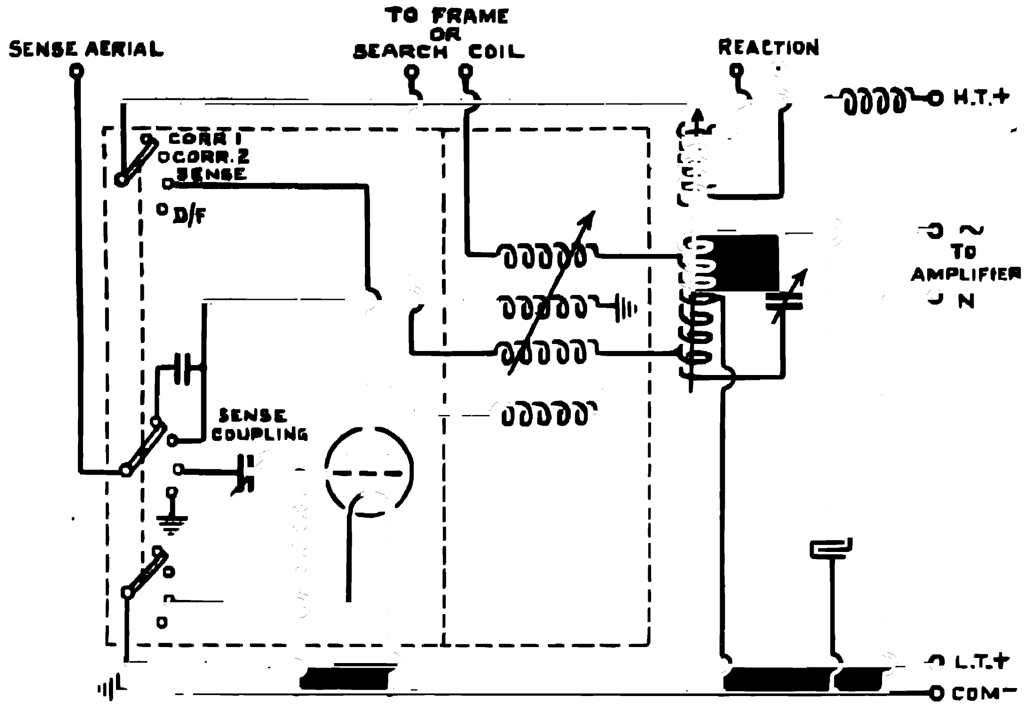

TO FRAME
OR
SEARCH COIL
SENSE AERIAL
REACTION
H.T.+
CORR 1
CORR. 2
SENSE
D/F
TO
AMPLIFIER
N
SENSE
COUPLING
L.T.+
COM−

FIG. 34.

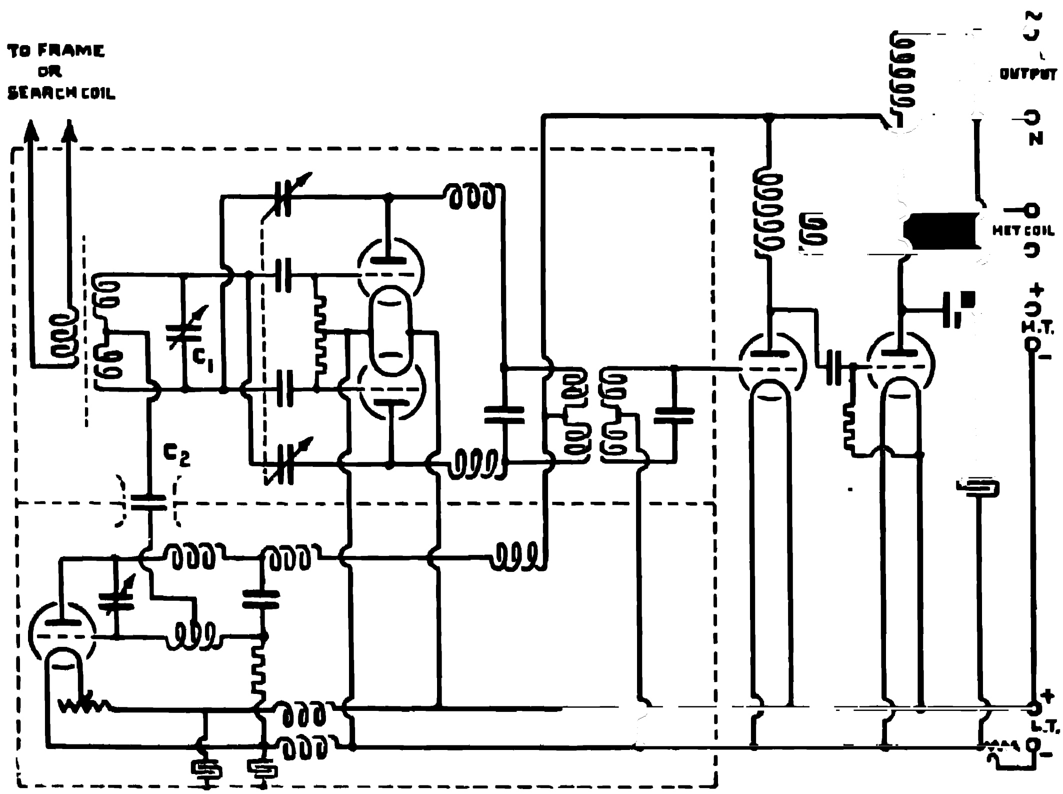

TO FRAME
OR
SEARCH COIL
OUTPUT
N
HET COIL
H.T.
L.T.
C1
C2

FIG. 35.

Operators are trained to search for signals on L/F and M/F in the "stand-by" position, with the sense aerial switch in the "sense" position; the search coil or frame coil need not be moved, for the vertical aerial is non-directional. With C.W. (Type A1 waves), the signal is picked up simply by manipulating the tuning condenser of the heterodyne oscillator. This method of searching will only fail in the unlikely event of the E.M.F. in the loop aerial or search coil being equal and opposite to that produced in it by the vertical sense aerial. After picking up a signal, the operator adjusts to the D/F position, and, finally, determines the sense of the bearing.

36. **Calibration of a Warship's Direction Finder.**—After installation, the outfit should be subjected to a schedule of tests to eliminate faulty gear; finally, it should be calibrated. The object of this process is :—

(*a*) To determine the necessary correction for quadrantal error on all frequencies. This may subsequently be applied by reference to a curve of corrections, or it may be applied automatically by a cam, a shunt inductance, or otherwise.

(*b*) To ascertain that any sense arrows or pointers are indicating correctly; to take note of the reliability of the sensefinder and the highest frequency at which it will work.

The whole operation calls for the highest technical skill. During calibration, the ship takes up a position 3 to 5 miles from a shore transmitting station but without intervening or adjacent land. The routine which is then followed depends upon the nature of the D/F outfit that is being calibrated. All bearings should be relative to the centre line of the ship and should not involve the use of the ship's compass.

Rotating Coils.

(*a*) The transmitting station is brought to bear right ahead. With the cam corrector at 0°, the pointer is set so that it reads 0° when a zero is obtained. It is usually preferable to isolate the ship's main aerial.

(*b*) The ship is then slowly turned through 360°, at not more than 6° a minute, in order to obtain a curve of corrections for various relative bearings of the transmitter, working at a given frequency. A direct comparison of the visual and wireless bearing is made at the rate of about 2 per minute, that is to say, about every 3° of swing. If the zeros are found to be blurred, use is made of a semi-circular corrector and a note is taken of the maximum coupling that has to be used. A record is made in the calibration report that if the amount of coupling is found to be in excess of a certain stated value on any subsequent occasion, it is necessary to suspect that there is a source or error in the instrument, or that night effect is present. Having obtained a curve of corrections, it is possible to arrive at a setting of the standard cam corrector—if this is fitted—which will automatically apply approximately the requisite correction on all relative bearings, for a given frequency. Alternatively, a specially cut cam (or cams) may be supplied.

(*c*) The ship is then steadied on red or green 45°, and on the quarter bearings if time permits, and observations are taken of the maximum quadrantal error for different representative frequencies. If unexpected values are obtained at any frequency, operation (*b*) is performed again. It is generally found that the curves of corrections for neighbouring frequencies conform in general shape to that on which the swing was performed. It is therefore only necessary to take note of the maximum quadrantal error on different frequencies, in order to be able to sketch the probable curves of corrections. The calibration report will contain a note of the settings of the standard cam corrector, to give approximate correction at each of the frequencies on which observations are taken.

(*d*) Finally, it is important to check the setting of any pointer which may be fitted to indicate "sense," and to test the functioning of the sensefinder.

Although correction for the greater portion of the deviation may be applied by a standard cam,

where the highest accuracy is required (*e.g.*, in Navigation), the cam corrector—if fitted—should be put out of action or neglected, and the correction applied from deviation tables or the curves of corrections.

Bellini-Tosi Systems.

(*a*) The aerials are approximately balanced by adjusting the series or shunt inductances which are provided for this purpose. To do this, the transmitting station is brought to bear successively green (starboard) 45°, green 135°, red (port) 45°, and red 135°, and adjustments are made to the quadrantal correctors until approximately no deviation is obtained at each of these points. This operation is performed on a number of representative frequencies. Shunt inductances are designed so that the bearing may be shifted at the rate of 1° per stop; the calibration report will contain a note of the setting of the inductance that gives good correction over a band of frequencies.

(*b*) The ship is then slowly swung through 360° in the manner already described for the rotating coil, in order to obtain a curve of residual corrections or deviation table. Where necessary, this operation is repeated for various frequencies. A cam corrector is sometimes fitted to the moving element of the goniometer, automatically to give approximate correction for the residual error over a band of frequencies. In some cases, it is necessary to supply two cams in order adequately to cover the frequency range. In this case when true bearings are read by means of the cam pointer, it is essential to be sure that the shunt inductance is on the stop appropriate to the frequency, as detailed in the calibration report.

(*c*) Finally, it is important to see that the sense arrow on the goniometer handle is set correctly and that the sensefinder functions properly.

Where bearings of the highest accuracy are desired, it is better to disregard the reading of the cam pointer, and to apply the necessary correction for residual error by direct reference to the curve of residual corrections.

37. **Direction Finders in Ships.**—From an electrical point of view, a direction finder should be situated as high above the hull of the ship as possible. This tends to reduce the amount of quadrantal deviation, but regard has to be paid to the position of aerials and other conductors such as main aerial halyards, lightning conductors, lead-cased cables for masthead lights, etc., all of which are capable of acting as vertical aerials. The currents in these conductors have the effect of blurring the zeros. Thus the best position for a direction finder is at the top of a mast, above the level of the roof of the main aerial and above all other conductors. When this is not possible, every endeavour should be made to obtain a position which is as symmetrical as possible, both about the centre line and athwartships, *e.g.*, Bellini-Tosi aerials between two equal vertical funnels are superior to aerials abaft a single funne. The same consideration should be borne in mind in the rig of the aerials themselves, so that they may be as symmetrical as possible with respect to surrounding structures.

In warships, of course, the position of the D/F aerials cannot be settled entirely by electrical considerations, but regard must be paid to questions of vulnerability, the arcs of fire of high angle guns, and other considerations. This sometimes leads to the aerials being at a considerable distance from the office. Technically this is bad, and every effort should be made to reduce the length of cable which has to be run from the aerials to the office. Even with the lowest capacity paper-insulated cables in use at the present time, every 100-ft. represents an effective capacity across the loop or frame of 0·5 jar, and a capacity to earth of 2 jars, which sets a serious limit to the highest frequency on which bearings can be taken.

Whether it is better to fit a rotating coil or a Bellini-Tosi system depends upon the circumstances which obtain in the particular ship under consideration. Where space is available, a Bellini-Tosi system, with large aerials on the centre line and approximately amidships, is recommended for use

at low frequencies; otherwise, it is better to use a rotating coil or crossed loops at the top of the mast. Where the office can be placed immediately under the aerials, it may be possible to take bearings on the higher frequencies (above 1,000 kc./s.). In general, it is possible to work more quickly with a Bellini-Tosi system than with a rotating coil, and this gives it an advantage for tactical purposes, and with a fading signal.

As the frequency increases, the errors due to the ship likewise increase rapidly in amount and complexity. Their correction becomes correspondingly more difficult, and in every ship installation there is a limiting frequency which varies with the class of ship and the nature and position of the direction finder, above which reliable D/F bearings cannot be obtained.

Best readings are obtained on the lower frequencies when other aerials are insulated. With the main aerial earthed, but far from resonance with the wave on which the direction finder is working, there is usually a slight blurring of the zero. This becomes worse as the aerial comes near to resonance, and at resonance it is possible for a bearing to be greatly in error, unless the D/F aerials are very well placed.

It is sometimes possible to have in a ship a single wire aerial working on the same frequency as that to which the direction finder is tuned, but it must be at a considerable distance in order to be safe. It is best to have it below the level of the direction finder, if possible.

It is important that calibrations should only be conducted under conditions likely to prevail in practice. It follows, therefore, that when using the direction finder for accurate work, care must be taken to see that those conditions have not been altered.

38. **The Adcock Direction Finder.**—Night effect errors have been shown to be due to the voltages induced in the horizontal parts of the loop. In 1916, Adcock suggested that the effect might be cut down by eliminating the horizontal parts, and designed a spaced aerial direction finder known as the "U" type. In 1929 Eckersley proposed a modification which will now be described and referred to as the screened "U" type, in which the lower horizontal leads to the receiver are screened by leading them underground through an earthed metal tube, so that no E.M.Fs. can possibly be induced in them by an incident wave, no matter what may be its state of polarisation. The loop direction finder is thus modified as shown in Fig. 36.

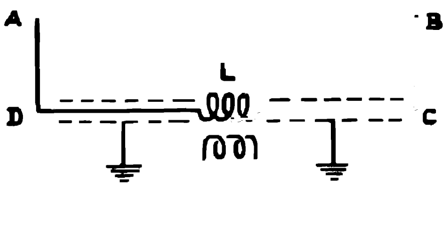


FIG. 36.

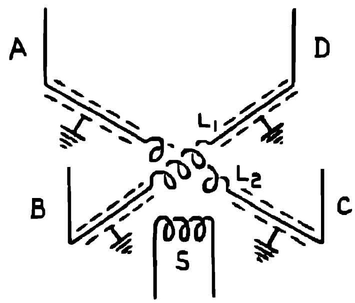


FIG. 37.

AD and BC are two vertical aerials corresponding to the vertical sides of a rotating loop, and are joined by an underground screened cable to an inductance L in the office, this inductance being coupled to the first stage of the receiver. The whole is to be considered capable of rotation about a vertical axis through its centre point. If the circuit is arranged symmetrically, the two aerials may be considered as earthed in common from the electrical mid-point of L. The currents flowing in opposite directions through the two halves of L from the vertical aerials to earth, are then in the same proportion as the E.M.Fs. induced in the aerials by the incident wave. The resultant flux through L is thus proportional to the difference of these two E.M.Fs., and the system is electrically equivalent to a loop aerial, except that it is only affected by the vertical component of the electric field of the wave. For example, when the plane of the aerials is at right angles to the direction of propagation of the wave, equal E.M.Fs. are induced in the aerials, equal currents flow in opposite

directions through the two halves of L and there is no resultant flux through L. Zero signal is therefore obtained in this position, as in the case of a loop aerial receiving the direct ray only. The complete theory of the effect of the screened tube has so far defined mathematical analysis.

The Adcock equivalent of the Bellini-Tosi system is shown in Fig. 37. Four equal vertical aerials, A, B, C and D, are situated at the corners of a square. Opposite aerials AC, BD are joined by means of buried cables through two equal inductances L_1 and L_2, mounted at right angles as in an ordinary goniometer. The search coil S rotates and is joined to a receiver suitable for the frequencies in use. The practical operation of such a type of Adcock system is exactly the same as a Bellini-Tosi system. In general, however, there is an error which is **octantal** in nature and will be described below.

The principle of the "Spaced direction finder" is found in use in many different forms. Another of these which gives good results employs the elevated or "H" type of aerial system. In fact it consists of two Hertzian half-wave length aerials, often called dipoles or doublets, and is shown in Fig. 38.

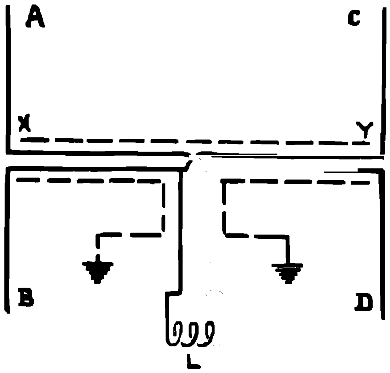


FIG. 38.

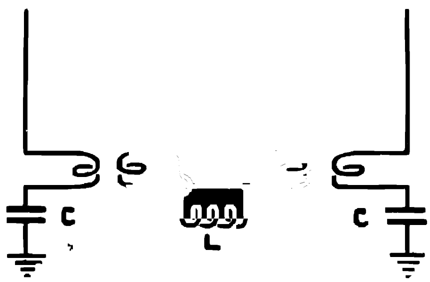


FIG. 39.

The Hertzian aerials AB, CD, are mounted vertically at the opposite ends of a horizontal tube XY, which can be rotated about a vertical axis through its centre point. The leads from the aerials are brought along inside the metal tube shown by the dotted lines. The current which flows through the coil L is then due to the difference of the E.M.Fs. induced in the two dipoles AB, CD. If the whole system is rotated until XY is perpendicular to the direction of the wave, no current flows through L, and a zero is obtained in the receiver. Here again, a Bellini-Tosi form of this type of Adcock aerial system is also possible.

In order to obtain satisfactory signal strengths it is desirable that the separation between the opposite aerials should be about one-tenth of a wavelength. At 300 kc./s. this means about 100 yards. The space required for Adcock aerials working on the medium and lower frequencies is therefore considerable. It should also be remembered that exact equality in the receiving power of opposite aerials is necessary if accurate bearings are to be obtained. It is clear that, even if sufficient space were available to permit the fitting of Adcock aerial systems in ships, this last requirement might be very difficult to secure. Any difference between the effect of the structure of the ship on each of the aerials would produce serious error. When Adcock aerials are used at high frequencies, extreme rigidity of design is essential. All four aerials should be accurately vertical to within a degree and of equal length to within an inch. Furthermore, it may be taken for granted that in favourable circumstances in a ship, one might get a separation between opposite aerials of about 10 yards; at a frequency of 300 kc./s., the differential action of the two aerials is about 6 per cent. of the response of either aerial alone. It has been calculated that to achieve an angular accuracy of 1° in these circumstances, the total extraneous pick-up must not exceed about 2 per cent. of this, which is 0·1 per cent. of the pick-up of either aerial alone. This represents an extremely

searching instrumental requirement. Finally, the inflexibility of this type of direction finder should be noted: an Adcock system giving reasonably good results at high frequencies, would be extremely insensitive at medium frequencies, and relatively much more so than a loop direction finder.

Another Adcock system, the advantages of which have recently been investigated by Barfield, is known as the "balanced coupled type," and is shown in Fig. 39 in its rotating loop form. The magnetic couplings aim at increasing the impedance between the vertical aerials and the horizontal feeders in order to reduce the current flowing into the vertical members, which is produced by the action of the horizontally polarised component of the electric field of the wave. The electrostatic screen between the windings produces this result. The vertically polarised component of the electric field produces its usual differential effect in L. The series condensers in the earthed leads are there to balance the paths to earth and so equalise the currents to earth. It appears that the residual error produced by night effect on a system of this kind may be reduced to a very small value; unfortunately, however, its sensitivity is very low.

In the range of the direct ray, and where the indirect ray is negligible, there is nothing to choose in point of accuracy between a closed loop direction finder and one based on the Adcock principle. Where the indirect ray is important, the Adcock has the advantage. Even this sytem will fail when a number of rays from the transmitter reach the receiver at the same instant from different directions. This appears sometimes to happen at high frequencies with a receiver which is outside the range of the direct ray, and is supposed to be due to the scattering of the rays in the outer atmosphere, like the scattering of the sun's rays by fog. At very great distances it appears that the main indirect stream of energy arrives in a fairly uniform direction in the great circle plane between the transmitter and the receiver and a definite bearing is again obtained.

39. **Octantal Error.**—The result obtained in paragraph 4, that the loop E.M.F. is proportional to the cosine of the angle between the plane of the loop and the direction of the transmitter, was

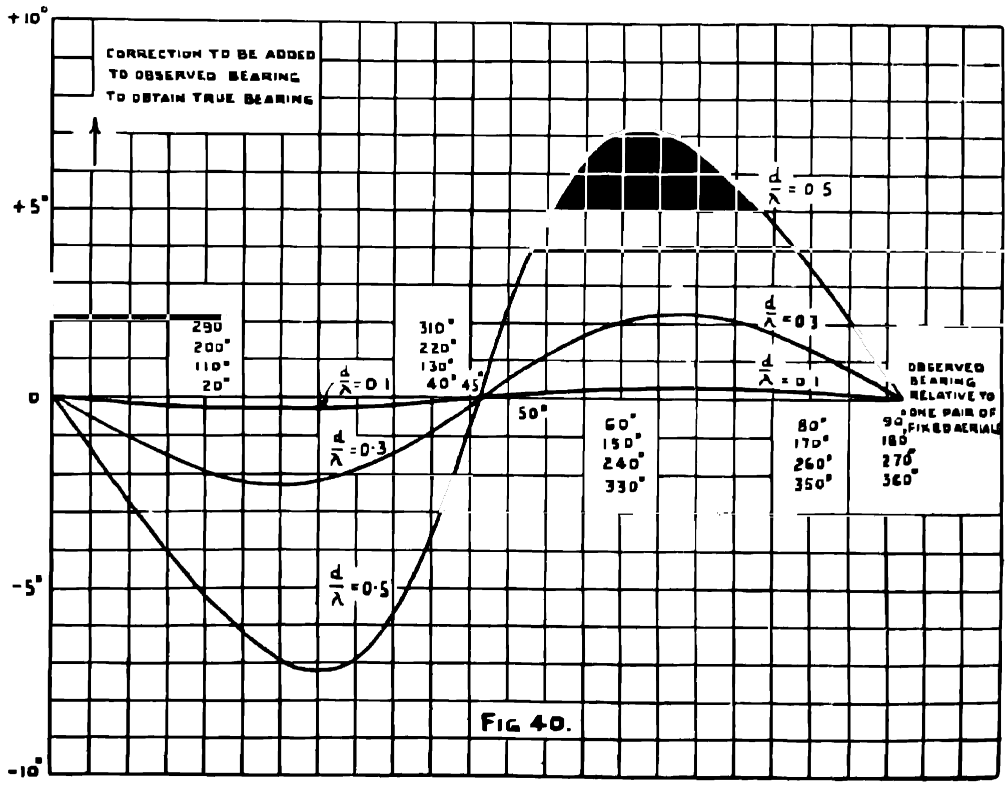


FIG. 40.

seen only to be justified provided that the width of the loop (or the distance between the two vertical aerials in the Adcock system) was small compared with the wavelength. At the higher frequencies at which these Adcock systems are most necessary, this condition may not be realised; particularly since their high impedance makes it necessary for them to be of large dimensions to give sufficient signal strength.

Provided the bearing is obtained by a zero on the rotating coil or its Adcock equivalent, this limitation does not affect the bearing, since the resultant E.M.F. is zero whatever the distance between the aerials in any practical arrangement.

In the Adcock equivalent of the Bellini-Tosi system, in general, the resultant magnetic flux in the goniometer is not always perpendicular to the direction of the incoming signal, as it was in the case analysed in paragraph 13, and the goniometer pointer will not give the true bearing. This will be the case unless the distance d between the vertical aerials (or sides of the loops) is small compared with the wavelength λ, in practice it should be less than about one-quarter of the wavelength. In fact this means that for incoming signals on all possible bearings from 0° to 360°, at the eight equally spaced **octantal points** 0°, 45°, 90°, etc., the resultant flux will be exactly at right angles to the direction of the signal and the pointer will indicate the true bearing. For signals received from all other directions, the bearing indicated by the pointer will be in error, and a correction must be applied. The magnitude of this correction varies with the ratio d/λ, and the direction of the wave. It is shown graphically in Fig. 40 for three values of d/λ, and the formula may be derived mathematically as shown below. This octantal error is not to be confused with another one, which is an instrumental one associated with goniometers in general, and not treated here.

It should be understood that this error is common to all Bellini-Tosi systems, but is of more importance in the Adcock equivalents which are usually employed at higher frequencies.

★40. **Mathematical Analysis.**—In Fig. 41, A, B, C and D represent the four spaced aerials, and a wave is shown incident at an angle θ to the plane of the pair CD. The wavefront is indicated, together with the projections from A, B, C and D. The semi width of the system is "a," the width being "d."

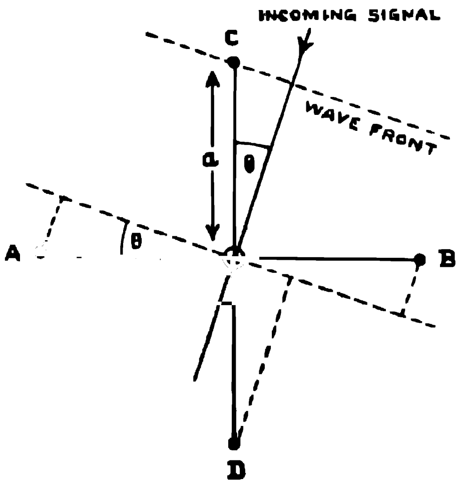


FIG. 41.

If the field at O be $\mathscr{X} \sin \omega t$

then that at C will be $\mathscr{X} \sin \left(\omega t + \dfrac{2\pi a \cos \theta}{\lambda}\right)$ (para. 4)

Similarly at D ... $\mathscr{X} \sin \left(\omega t - \dfrac{2\pi a \cos \theta}{\lambda}\right)$

and at A ... $\mathscr{X} \sin \left(\omega t - \dfrac{2\pi a \sin \theta}{\lambda}\right)$

and at B ... $\mathscr{X} \sin \left(\omega t + \dfrac{2\pi a \sin \theta}{\lambda}\right)$

If h is the height of the vertical aerials, then as in paragraph 4, by taking the difference between the instantaneous E.M.Fs. in the two members constituting a pair we get—

E.M.F. in pair CD $\quad 2\mathscr{X}h \sin \left(\dfrac{2\pi a \cos \theta}{\lambda}\right) \cos \omega t$

E.M.F. in pair AB $\quad 2\mathscr{X}h \sin \left(\dfrac{2\pi a \sin \theta}{\lambda}\right) \cos \omega t.$

With identical impedances, currents, and hence magnetic fluxes, are produced in the goniometer windings which are proportional to the E.M.Fs. (*cf.* paragraph 13). Further since the E.M.F. in AB produces a flux in the plane of CD, and that in CD produces one in the plane of AB, if α be the angle which the direction of the resultant flux makes with the plane of AB, then we have

$$\tan \alpha = \frac{\sin \left(\dfrac{2\pi a \sin \theta}{\lambda}\right)}{\sin \left(\dfrac{2\pi a \cos \theta}{\lambda}\right)} \quad \ldots\ldots\ldots\ldots \quad (1)$$

From (1) it may be seen that α and θ will be coincident in value at the octanal points and the true bearing of the wave will be given.

For we have, when $\theta = 0°$, $\sin\theta = 0$, and $\alpha = 0° = \theta$,
when $\theta = 45°$, $\sin\theta = \cos\theta$, $\tan\alpha = 1$, and $\alpha = 45° = \theta$,
when $\theta = 90°$, $\cos\theta = 0$, $\tan\alpha = \infty$, and $\alpha = 90° = \theta$,

and so on. Between these points, the maximum value of the difference between α and θ will depend upon the ratio $\frac{d}{\lambda}$, and on θ, in the way shown in Fig. 40. The mathematical analysis for this will not be given. Finally, from (1) it is clear, that when $\frac{d}{\lambda}$ is sufficiently small, the sine of the angle is replaceable by the angle itself in circular measure, and $\tan\alpha = \tan\theta$, that is $\alpha = \theta$ always.

41. **Directional Transmission : Radio Beacons.**—The principle of the wireless lighthouse is in use in various parts of the world in three forms :—

(*a*) Flashing radio beacons, with all round radiation.
(*b*) Revolving radio beacons.
(*c*) Radio-range beacons.

(*a*) Flashing radio beacons are in all respects like omni-directional flashing lighthouses, and used in the same way. A characteristic morse signal radiated at stated intervals takes the place of the flashings of lighthouses and makes recognition easy. They usually work on some frequency in the neighbourhood of 300 kc./s., and operate continuously in fog and at specified times or on request. For ships, or aircraft, having direction finding apparatus, it is generally better to use these beacons for D/F work and not commercial or other wireless stations, unless the geographical position of the transmitter is known with accuracy and the bearings are likely to be free from error due to coastal effect or other causes.

(*b*) Revolving beacons render possible the determination of direction or bearing from a fixed point without the use of special direction finding apparatus in the ship or aircraft. Quite simple receiving apparatus is required to observe either the instant at which a sharply focussed beam sweeps past the sector that includes the receiving ship, or to detect a rise and fall in signal strength as the beam rotates. This is "directional transmission" in the sense in which this term was used in paragraph 1.

The methods used in the production of revolving beacons depend upon the frequency to be employed. Employing frequencies of the order of 30,000 kc./s. it is possible to design narrow uni-directional beam transmitters, in which the directional effect is obtained by having some reflector system which rotates around the radiating aerial. One of the first of such beacons was installed at Inchkeith by the Marconi Company, and had a single wire vertical aerial transmitter, with a parabolic reflector which could be rotated about this aerial. The reflector consisted of a number of vertical wires in the form of a parabola, with the transmitting aerial at the focus. The beam could be made to revolve by rotating the reflector system about the transmitting aerial as axis. Although the width or aperture of the parabola was only about two wavelengths, the beam was sufficiently sharp to give an angular accuracy of 2°-3° at a range of ten miles. Its working range was, however, rather restricted, and it had the disadvantage that a ship making use of the beam needed a special receiver suited to the very high frequency in use. Beams working on frequencies of the order of 80,000 kc./s. have been used in America for assisting aircraft to make good landings in fog.

At frequencies in the medium frequency range use may be made of the directional properties of a loop aerial. A loop employed as a transmitter will have a horizontal polar diagram in the form of a figure-of-eight. The maximum radiation is in the plane of the loop, while in the plane at right

angles to the loop there is zero radiation. Thus, if such a coil is made to rotate clockwise through 360°, say once a minute, the direction of zero radiation also turns through 360°. Now suppose that the plane of the coil is due East and West when the revolution commences, and a signal is sent out, such as the letter " N," to indicate the exact instant at which the coil starts to rotate. Imagine an observer at a point bearing 135° (true) from the rotating coil receiving signals on an ordinary open aerial. At the instant at which he hears the letter " N " he starts a stop watch. The signal gradually increases in strength until, after $7\frac{1}{2}$ seconds, it reaches a maximum. It then commences to decrease, and at the end of $22\frac{1}{2}$ seconds the rotating coil is at right angles to the direction of the receiver, and a zero is obtained. The signal again commences to increase, reaches a maximum, and once more returns to zero after a total time of $52\frac{1}{2}$ seconds from the commencement of the revolution. It is clear that by taking the time at which the zero is obtained in the receiver, the bearing of the transmitter can be obtained. If the zero occurs t seconds after the commencement of the swing, then the bearing is

$$6t^\circ \quad \text{or} \quad (6t + 180)^\circ,$$

since the rate of rotation is 6° per second.

It is only possible to say which of the bearings is the correct one, provided sufficient is known of the probable direction of the transmitter, for at present no arrangements are made to remove the 180° ambiguity. In case the letter " N " should not be heard because the receiver is due North or South of the beacon, a second signal is sent out when the plane of the coil is North and South.

This system of directional transmission, in which there is some radiation in every direction but one, has the great advantage that it can be used at medium frequencies, and demands no special apparatus at the receiving end. An ordinary open aerial and receiver is all that is necessary, together with a stop watch. As the time of the revolution of the beacon is maintained very accurately at 60 seconds, it is possible to check the rate of the stop watch. This should always be done and allowed for in the calculation of the bearing. For example, if the zero is obtained 20 seconds after the commencement of a revolution, but the time of revolution appears to be 61 seconds by the stop watch, the true bearing is not 120°, but

$$\frac{20}{61} \times 360 = 118^\circ.$$

In order to get sufficient radiation from a coil of moderate dimensions it is necessary that it should carry very large currents. For example, one beacon at present in use has four turns on a frame 10 ft. × 11 ft., and carries a current of 80 amperes, at a frequency of 300 kc./s. Instead of a rotating coil, two large loops at right angles can be coupled to a transmitting goniometer.

All forms of revolving loop transmitter are liable to " night effect," since in general it is not possible to prevent the radiation of an electric field from the horizontal parts of the loop and which is horizontally polarised. Theoretically, therefore, an improvement should be obtained by using the transmitting equivalent of the Adcock receiver, and this is sometimes employed with the radio-range beacons. From this it follows that aircraft employing a trailing aerial cannot hope for accurate results.

(*c*) Radio-range beacons, employing medium frequencies, are fixed in direction, and are used in some parts of the world to define safe courses for aircraft, or to assist in homing. These beacons are produced in various ways and several methods are used to indicate the course to be followed. Broadly, there is the type of beacon that gives an aural indication of the course, and another that gives a visual indication on an indicator like a frequency meter. In the first type two crossed loops may be used, each loop being alternately energised from a common source. The resulting radiated energy will be in the form of two figures-of-eight as in Fig. 42.

In one such system the letter A is transmitted from one loop and the complementary letter N from the other loop, the signals being so timed that with respect to a receiver situated somewhere on the equi-signal line, they coalesce and produce a continuous dash. So long as the aircraft

remains on the course the dash is heard. If it moves off the course a preponderance of A or N will indicate the direction of the error.

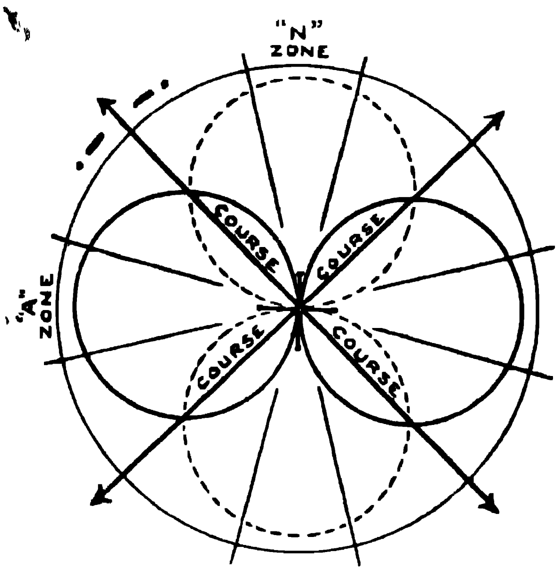


FIG. 42.

In the second type employing visual indication, a simple method is to excite the two loops in the same phase, but to employ a carrier having a different modulating frequency in each case. The two modulating frequencies might be of the order of 65 and 86 cycles per second. After detection, these two frequencies may be arranged to operate the reeds of a frequency meter. So long as the aeroplane moves along the equisignal course, the amplitude of vibration of the two reeds will be equal. If this course is lost the reeds will vibrate with unequal amplitude and the inequality will point out the direction of the error. Fig. 43 shows the directional characteristics produced in this way, and it can be shown that only two courses result, as opposed to the four courses produced by the aural type. Four courses can, however, be produced, and the angle between them can be controlled by altering the phase equality of the two loops, or by making the percentage modulation in the two cases unequal.

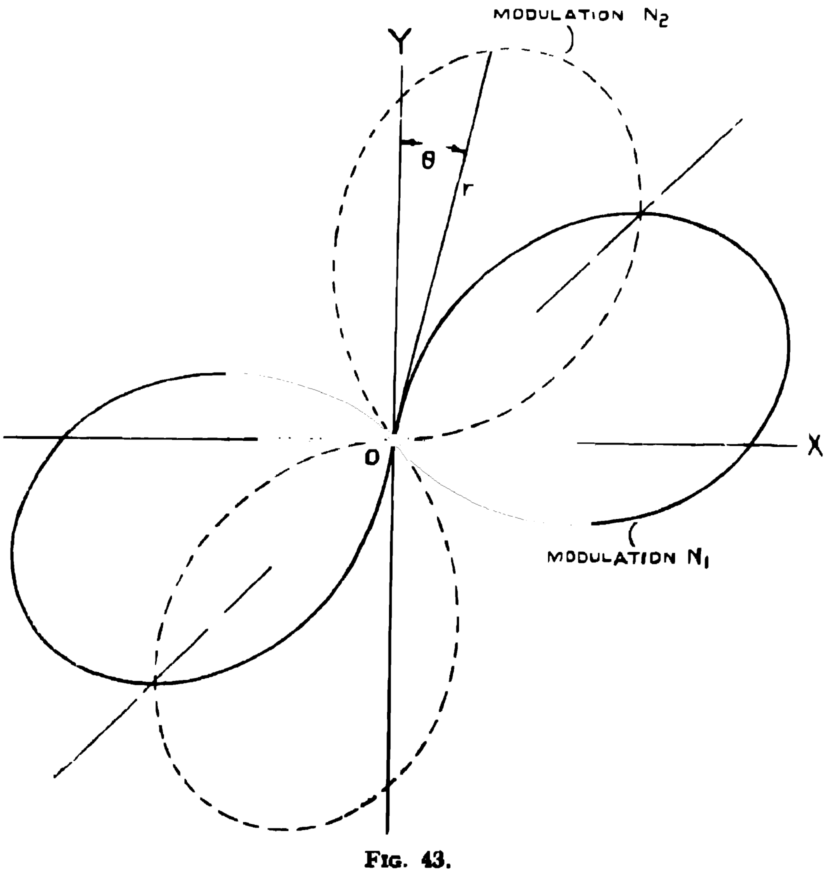


FIG. 43.

High-frequency radio-range beacons are in use in different parts of the world to facilitate the blind landing of aeroplanes in fog. In the Lorenz System, the pilot follows the homing beacon at a steady altitude; at about $1\frac{3}{4}$ miles from the landing position, the machine runs through a vertical wall of signals, some 200 yards in width in the direction of flight; at this point the pilot begins to glide to earth following a line of equal signal strength. Further on, the machine penetrates another vertical wall of signals, this one about $\frac{1}{4}$ mile from the point of contact with the ground; the pilot begins to "flatten out." In the machine, three separate visual receivers are provided, two of these to receive the two warning signals, and the other to observe the signal strength of the beacon signals. The visual indicators may conveniently be neon lamps. Arrangements are also made to provide acoustic reception of the two warning signals, and of the main signal from the radio beacon. Beacons of this kind use frequencies of the order of 33·3 megacycles, and the two warning signals use a frequency of the order of 38·0 megacycles. The two warning signals, in the form of vertical walls, are produced by the use of horizontal dipoles and reflectors, a different modulating frequency serving to distinguish the first warning signal from the second.

The adjustment of the beacon to provide a suitable gliding angle when the pilot is following a line of equal signal strength, depends upon a nice control of the vertical polar diagram of the beacon. In practice the angle of maximum radiation is inclined upwards at an angle of about $8°$ to the horizontal, and a line of equal signal strength may be regarded as a curve within the envelope of the vertical polar diagram which is defined by the requirement that the rate of **decrease** of signal strength due to loss of angular height must be made equal to the rate of **increase** of signal strength due to approach to the beacon.

EXAMINATION QUESTIONS IN D/F.

Numerical Examples.

1. Find the voltage induced by a plane wave of field strength 0·01 volts per metre and frequency 1,000 kc./s. in :—

    (*a*) A vertical aerial 8 metres high.

    (*b*) A frame aerial 1 metre square of 12 turns, the plane of the frame being in the plane of propagation of the wave.

    [Result: (*a*) 80 mV.; (*b*) 2·51 mV., L.U., 1932.]

2. A frame aerial has 20 turns of diameter 6-ft., the resistance of the coil being 2·5 ohms at 100 kc./s. If a signal of this frequency has a field strength of 30 micro-volts per metre, what current will it produce in the coil when it is tuned to the signal?

    (1·32 micro-amps., I.E.E., 1926.)

3. A frame aerial consists of 40 turns; each turn is a square of 1·5 metres side, and the whole has an inductance of 9,000 micro-henries. The frame is tuned by a variable condenser. If the frame aerial has a resistance of 8 ohms, what will be the voltage across the condenser when the frame and condenser are tuned to resonance with an incoming field of 20,000 micro-volts per metre at a frequency of 60 kc./s.?

    (0·959 V., C. & G.; Final, 1929.)

SECTION "T."

4. A loop of area 1 sq. metre, and having 10 turns has its plane at 45° off the line of bearing of a distant station which produces an incident magnetic field of strength 10^{-9} lines per sq. cm. at a frequency of 300 kc /s. Find the E.M.F. induced in the loop.

   (1·33 micro-volts, I.E.E., 1930.)

5. A small frame aerial is connected across a calibrated variable condenser. A valve voltmeter is also connected across the condenser. The frame is placed so that its plane is inclined at 45° to the direction of a distant transmitting station sending a continuous dash on 100,000 cycles per second.

   Maximum deflection of the voltmeter is obtained when the condenser is adjusted to 1,000 micro-micro-farads. The frame is then turned until its plane is in the direction of the transmitting station and the condenser adjusted to give an equal deflection on the valve voltmeter. The capacity of the condenser is then 990 micro-micro-farads. What is the inductance and resistance of the frame aerial circuit ?)

   (Result : 2·53 mH. 15·6 ohms. C. & G Final, 1933.

Descriptive Examples.

1. Account for the difficulty of determining the direction of H/F signals. What means can be adopted (*a*) on shore stations, and (*b*) in ships, to minimise the difficulty ? Enumerate the various errors to which a D/F system is exposed, and briefly state the means adopted for counteracting them in a Bellini-Tosi system.

   (Qual. Wt. Tel., 1935.)

2. Describe a frame aerial. What are the properties which distinguish this type of aerial from a simple vertical type of aerial ?

   Describe the theoretical basis of sense finding in D/F work, and show how sense is found with a Service type rotating loop outfit.

   (Qual. Lts. (S), 1935.)

3. Show briefly how a Rotating Frame Aerial can be used to determine the direction of transmission. State how sense is determined and explain how the necessary conditions for obtaining a good sense bearing are fulfilled. Illustrate your answer by polar diagrams.

   (W/T.1., 1934.)

4. Describe a direction finding system suitable for use in ships. Give sketches showing the arrangement and screening of the coil and discuss the principles involved. Outline the rotating beacon system used for giving a bearing to ships at sea.

   (I.E.E., 1932.)

5. What is the cause of night errors in rotating loop and Bellini-Tosi direction finders ? Explain clearly how the Adcock aerial tends to obviate such errors. Give diagrams of the aerials and associated circuits for (*a*) rotating Adcock system, and (*b*) fixed Adcock system.

   (C. & G. Final, 1934.)

R/F MEASUREMENT—WAVEMETERS AND OSCILLATORS.

1. **Importance of Radio Frequency Measurement.**—The operation of receivers and transmitters, under modern conditions, involves the following practical problems :—

(*a*) Measurement of the frequency of an incoming signal; it may come from a distant station, or it may be injected from a local transmitter which is to be set to a "spot frequency."

(*b*) Tuning a receiver to a given frequency.

The requirement in (*a*) involves the use of a piece of apparatus still generally called a **wavemeter,** when it is a precision instrument, or a **wave indicator** when the determination is less exact. A device giving an indication on one frequency only, is a familiar feature in some transmitters working on fixed frequencies; it is commonly called a **frequency monitor.**

The requirement in (*b*) involves a method of energising the tuned circuits of a receiver. This needs a **local oscillator**; its tuned circuits may be directly calibrated and, for portable use, this will probably be convenient. Since the calibration varies when different valves or battery potentials are used, such an instrument can never be capable of the highest accuracy. It is usually preferable to have a roughly calibrated local oscillator and to set it to its required frequency by working it in conjunction with a wavemeter. In somewhat earlier days, the name **wave tester** was frequently applied to the local oscillator.

It is proposed to consider briefly the theoretical principles underlying the various types of wavemeters, wave indicators, and oscillators; it will be seen that in some respects their functions are interchangeable.

2. **Demand for Increased Accuracy.**—The greater frequency stability of modern transmitters has led to a greater demand for accuracy in frequency determination. For this reason, one sees the gradual relegation of older "wavemeters" to the status of "wave indicators," though, possibly, the first name still lingers. Apart from interference problems, a practical figure for the accuracy required in the tuning of transmitters and receivers is that which ensures reception of the signal; that is to say, a receiver anticipating a C.W. signal and using the heterodyne principle of signal detection, must be tuned ready for reception and not differ from the signal frequency by more than an audible amount. Modern instruments with superheterodyne circuits, etc., provide a selectivity demanding no greater difference than 2 kc./s. on the higher frequencies. Since two independent tuning processes are involved, one at the transmitter and the other at the receiving station, the frequency adjustments at each must be made to within $\pm$ 1 kc./s., in order that the operator at the receiving station may instantly pick up the signal without undue searching.

With suitable precision apparatus, frequency determination presents no difficulty to the laboratory worker, but laboratory apparatus, besides being expensive, requires skilled attention, and conditions of use that are an impossible proposition on board ship and in remote shore stations. Instruments for Service use are required to combine simplicity of operation, some degree of absolute accuracy, and a form that is, relatively, compact and robust.

Portable wavemeters of the simpler "absorption type," or wavemeters which are required to cover a very large frequency range, cannot possess the requisite degree of accuracy which is now necessary in order that transmitters may be set to their frequencies, within the internationally agreed limits of frequency tolerance. Such instruments have a guaranteed accuracy which never reaches one part in 1,000, and allowing for temperature and other effects, it is more likely to be of the order of one part in 250.

For the above reason, it is now common practice to include a wavemeter possessing a very high degree of accuracy, as an integral part of every modern transmitter. It is only in this way that the interference troubles of an overcrowded æther can be prevented.

3. Principle of Absorption Wavemeters.—These consist, essentially, of a simple oscillatory circuit, in which either the inductance or the capacity can be varied to cover a range of frequencies together with some device for indicating when the current in the circuit is a maximum. Such wavemeters absorb a certain amount of power from the source; hence their name. In the Service, the title " absorption " is usually dropped and these instruments are simply called " wavemeters." They are suitable for measuring the frequency of medium power and high power transmitters, and of incoming distant signals with the help of a suitable receiver.

If an alternating E.M.F. of constant amplitude is applied to a circuit consisting of inductance, capacity, and resistance in series, then the current has its maximum value when the circuit is tuned to resonance with the applied E.M.F. The resonant frequency " f " is connected with the inductance " L " and the capacity " C " by the well-known formula

$$= \frac{1}{2\pi\sqrt{LC}} \qquad \ldots\ldots\ldots\ldots\ldots\ldots \qquad (1)$$

With the quantities in (1) expressed in particular units, f in kc./s., C in jars, and L in mics., the formula becomes $f = \dfrac{3 \times 10^4}{2\pi\sqrt{LC}}$.

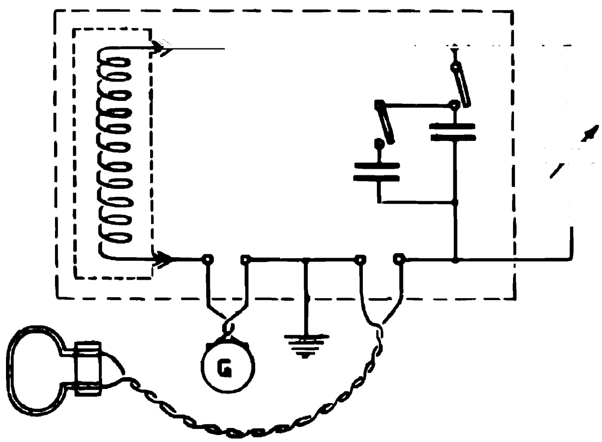


FIG. 1.

Fig. 1 represents a very old Service wavemeter, consisting of a closed oscillatory circuit which has, in series with it, a device for indicating when the current in the circuit is a maximum.

In general, according to its manner of operation, the indicating device may be either in series or parallel with the oscillatory circuit; various devices will be discussed shortly. The E.M.F. may be obtained either by loose magnetic coupling between the wavemeter circuit and the oscillator whose frequency it is desired to measure, or by loose capacitive coupling; the latter is now the more common. It will be remembered that if two oscillatory circuits are tightly coupled together a double frequency effect is produced; for wavemeter work it is, therefore, important that the coupling should be loose. Moreover, with tight coupling, the large currents produced may damage the indicating device.

The frequency at resonance may be read directly from a scale attached to the instrument, or from the calibration curves issued with the wavemeter.

4. The Components of the Tuned Circuit.—Special attention must be given to the design of the condensers and inductances constituting the tuned circuits. Tuning condensers are generally

designed so that there is a simple relation, called the law of the condenser, between the condenser setting and the frequency. Fig. 2 illustrates the following four cases that occur in practice :—

(*a*) Condenser setting proportional to the capacity. $(\theta \propto C.)$

(*b*) Condenser setting proportional to the wavelength. $(\theta \propto \lambda.)$

(*c*) Condenser setting proportional to the frequency. $(\theta \propto f.)$

(*d*) Condenser setting and frequency related by a logarithmic law of the form $(f \propto e^{k\theta})$.

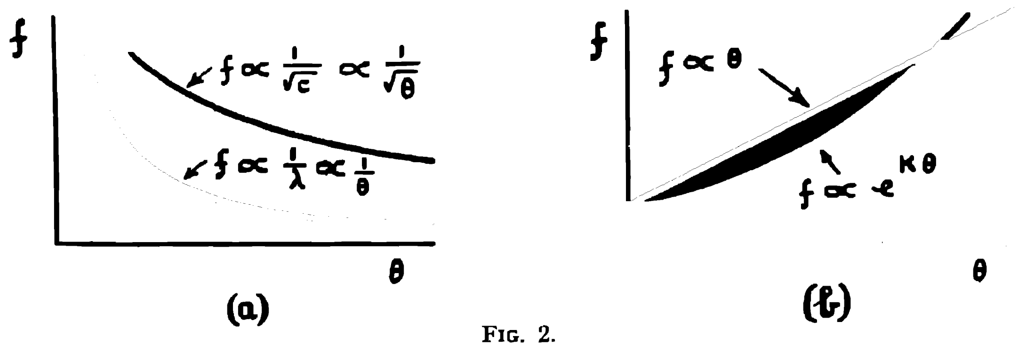


FIG. 2.

(*a*) This is the "straight line capacity condenser," and is the case of all of the older condensers with semi-circular plates; equal changes in angular overlap of the fixed and moving plates give equal changes in area and in capacity. The LC value of the circuit alters in the same proportion.

From (1) para. 3, we have $f \propto \frac{1}{\sqrt{C}}$, and since $\theta \propto C$

$$f \propto \frac{1}{\sqrt{\theta}}.$$

This means that if the frequency is to be halved, the reading must be four times greater ; a change in reading from 20 to 80 would produce the same frequency change as that from 2 to 8. It is clear that a law of this kind leads to a wide separation of the frequencies at the L/F end, and relatively crowds them at the H/F end, and interpolation is rendered difficult. A condenser of this form is now rarely used.

(*b*) A "straight line wavelength condenser" sometimes has certain advantages, and was very much used in the early days of broadcasting work when all stations were referred to in terms of their wavelength. In this case we have—

$$\theta \propto \lambda \propto \frac{1}{f} \propto \sqrt{C}.$$

Hence

$$C \propto \theta^2.$$

The design of the condenser plates is based on the above relation, the form of which led to the general use of the name "square law condenser."

In this case, if the condenser setting is doubled, the capacity becomes four times greater, the wavelength is doubled and the frequency is halved ; the relative frequency crowding at the H/F end is even worse. This condenser is, also, now somewhat out of fashion.

(*c*) Since broadcasting stations are now spaced on a frequency basis, with carrier waves 9 kc./s. apart, it is evident that it became more convenient to design condensers so that one division on the scale always represented the same alteration of frequency. For example, a condenser with plates capable of moving 180° usually has a scale with 100 divisions, giving 101 points ; for some purposes it might be convenient to arrange that a movement of one scale division corresponded to an alteration in frequency of 9 kc./s The tuning range would include 101 different stations, and could, for example, cover the range of frequencies from 600 to 1,500 kc./s.

Moreover, these straight line frequency condensers are almost the ideal ones for use in certain wavemeters. In this case—

$$\theta \propto f \propto \frac{1}{\sqrt{C}} \propto \frac{1}{\lambda}$$

The first condenser plates designed according to this law were long and narrow and resembled butterfly wings in shape. In that form they were somewhat bulky. Considerable improvement in design has since taken place and it is now possible to produce a relatively compact condenser following the above law.

(*d*) It is possible to approximate to straight line conditions by designing a condenser that follows a logarithmic law. It is obvious that the "accuracy" at the low capacity end of the scale will be much increased if the rate of the change of capacity with scale angle is simultaneously low, and that the accuracy at the other end will not be seriously impaired if the rate of change is there proportionately higher. This gives a possible design criterion that the "rate of change of capacity shall always be proportional to the capacity." This will produce equality of percentage frequency alteration $\left(\frac{\delta f}{f}\right)$ for equal alterations of condenser setting ($\delta\theta$) over the whole scale, a state of affairs which makes the best use of the smallest scale divisions. It is equivalent to the production of a condenser having a constant percentage error in frequency measurement over the whole scale. In terms of the first of these criteria, mathematicians will see that this means

$$\frac{dC}{d\theta} = -KC \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

where K is a constant.

The negative sign implies that C is arranged to decrease as θ increases, so making an inverted scale in order that f may increase with increase of θ. From (1), by integration, we obtain the expression for C in terms of θ in the form

$$C_\theta = C_{max}\, e^{-K\theta},$$

where C_θ is the capacity at any scale angle θ, where C_{max} is the maximum capacity, where K is a constant, and where e is the base of natural logarithms.

It will be seen that when $\theta = 0$ then $C_\theta = C_{max}$.

Since $f \propto \frac{1}{\sqrt{C}}$ .... it follows that

$f \propto e^{k\theta}$ .... where k is a constant.

Fig. 2 (*b*) shows the graph of $e^{k\theta}$, and it will be seen that, over a limited portion of it, the curvature is not very pronounced; for this reason the name "mid-line" condenser is sometimes used.

The absolute error in frequency will always be greater at the H/F end than at the L/F (or large capacity) end of the scale. Extreme variation may be countered by arranging that the ratio of maximum capacity to minimum capacity is not too great; in one Service example it is fixed at 5·6 : 1 (0·5 jar to 0·09 jar), giving a frequency ratio of a little over 2 : 1. In design work it is usually necessary carefully to arrange that the **minimum** tuning capacity in parallel with the inductance is relatively large; partly for this reason a fixed condenser will often be found in parallel with the variable one.

The condenser is, generally, the variable component because of the difficulty of constructing variable inductances which will preserve their values unchanged. This is mainly due to "dead end" effects.

When a wavemeter is required to cover a large range of frequencies it is usually supplied with a number of interchangeable coils; the inductances are selected so that the LC value obtained with any one coil and a high value of the variable condenser, may also be obtained with the next coil in the series and a low value of the variable condenser. In this way, a continuous band of frequencies is covered with suitable overlap on each range.

The self-capacity of the coils is taken account of in the calibration of the wavemeter, and care is taken that the resonant frequency of the coil itself is considerably above the range of frequencies over which it is to be used.

★5. **Mathematical Note.**—For a single condenser in parallel with a fixed inductance we have

$$\propto \frac{1}{\sqrt{C}} \text{ or } f = KC^{-\frac{1}{2}}$$

$$\frac{df}{dC} = -\tfrac{1}{2} KC^{-3/2} \text{ .... or } \delta f = -\tfrac{1}{2} KC^{-3/2} . \delta C = -\tfrac{1}{2} f/C . \delta C.$$

is the error in f corresponding to an error δC in the capacity C. The percentage error is given by

$$\frac{\delta f}{f} = -\tfrac{1}{2}\frac{\delta C.}{C} \quad \text{.................................} \quad (2)$$

This means that a given percentage error in the capacity produces half of that percentage error in the frequency. If the error in capacity be assessed at one part in 1,000, the maximum error in frequency introduced would never exceed one part in 2,000.

From (1) of paragraph 4, and (2) we have—

$$\frac{\delta f}{f} = K_1 \delta\theta. \quad \text{..............................} \quad (3)$$

This result makes clear the last of the criteria referred to in (*d*)—paragraph 4—above; the percentage error $\frac{\delta f}{f} \times 100$ is constant, over the whole scale range.

6. **Indicating Devices.**—The device for indicating when the current in a resonant or wavemeter circuit is a maximum may take several forms. The commonest are:—

(*a*) A pea lamp.
(*b*) A neon tube.
(*c*) A thermo-junction and galvanometer.
(*d*) A two-electrode valve and galvanometer.
(*e*) A triode valve and milliammeter.
(*f*) A luminous quartz resonator.

(*a*) A low voltage "pea lamp" may be inserted in series in the wavemeter circuit, or in a separate circuit mutually coupled to the main wavemeter circuit. The coupling of the wavemeter to the oscillating circuit under test should be so adjusted that the lamp just glows when the wavemeter current is a maximum. Sometimes a battery is inserted in the circuit to bring the lamp very nearly to the glowing point, so that only a small current and loose coupling are necessary to bring it to the indicating point. Fig. 3 shows a simple wavemeter circuit, with a mutually coupled lamp as indicating device.

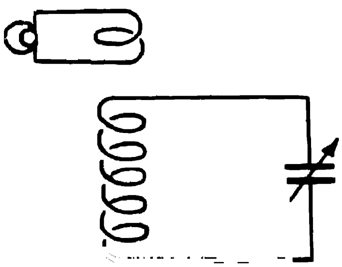


FIG. 3.

(*b*) The neon tube is connected in parallel in the wavemeter circuit, *i.e.*, across the condenser. It depends for its action on the fact that, when sufficient potential difference is impressed across a tube containing neon gas at low pressure, ionisation occurs, and the gas glows. When maximum current flows in the wavemeter circuit, the potential difference across the condenser is a maximum, if the resistance in the circuit is negligible, and, hence, the neon tube serves as an indicator of maximum current.

Fig. 4 shows a wavemeter circuit using a neon tube as indicating device. The variable component in this circuit is the inductance, which is a variometer, and so avoids the difficulty of variable tapping points. The range is altered by switching in different fixed condensers.

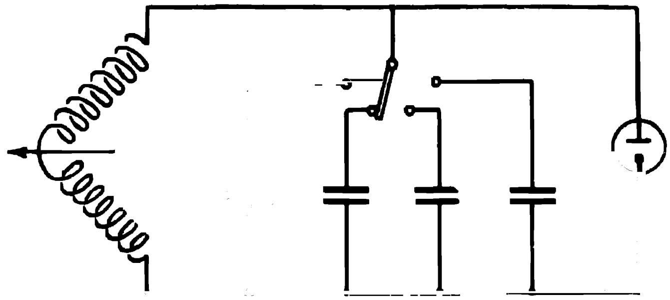
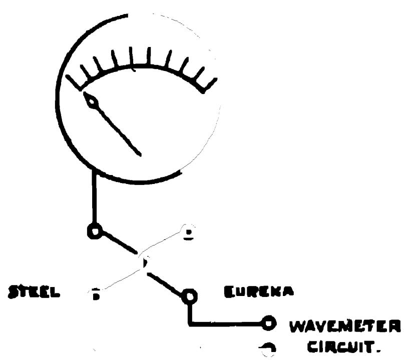


FIG. 4

(c) Thermo-junction and Galvanometer. If two wires of certain dissimilar metals are joined together, it is found that an E.M.F. is developed across the junction. This is probably due to the different concentration of the free electrons in the two metals, which tries to equalise itself across the junction. If the other ends of the metal wires are also joined so that a complete electrical circuit is formed, an equal E.M.F. will be developed at the second junction ; but as the two E.M.Fs. are in opposite directions round the circuit, there will be no resultant E.M.F., and no current will flow round the circuit. The E.M.F. produced in this way, however, depends on the temperature of the junction, and, within certain limits of temperature, the size of the E.M.F. is found to be roughly proportional to the temperature. If one of the two junctions is now heated so that it is kept at a higher temperature than the other, the E.M.F. developed at the hotter junction is greater than that at at the colder junction, there will be a resultant E.M.F., and a current will flow round the circuit. This is called the " Seebeck Effect," after its discoverer.

It is found that the resultant E.M.F. is unaffected by including a wire of some other metal as part of the circuit, and so a galvanometer, for instance, may be inserted in the circuit without altering the value of the resultant E.M.F. The extra resistance will, of course, alter the value of the current produced, and, as the thermo-electric E.M.F. is itself small, the galvanometer which indicates the size of the current should have a low resistance.

The utilisation of this effect in an indicating device is shown in Fig. 5. Two fine wires of dissimilar metals are interlinked, and lightly soldered at the junction. The metals steel and Eureka were originally used, but copper and Eureka give a greater E.M.F., an added advantage being that the copper does not rust. One pair of ends of the two wires is joined through a sensitive galvanometer, and the thermo-junction is connected in series with the wavemeter circuit by means of the other pair.



THERMO-JUNCTION AND GALVANOMETER.
FIG. 5.

The necessary rise in temperature of the thermo-junction is provided by the heating effect of the radio-frequency current flowing round the wavemeter circuit. The heating effect is proportional to the square of the current, and so the rise in temperature of the junction is greatest when the current in the wavemeter circuit has its maximum value, *i.e.*, when the wavemeter circuit is tuned to resonance with the oscillatory circuit whose frequency it is desired to measure. In consequence, the reading of the galvanometer is greatest at this setting of the wavemeter condenser.

The current through the galvanometer is a direct current, due to the unidirectional thermo-electric E.M.F. produced at the hot junction. It should be clearly understood that no radio-frequency current flows in the galvanometer which, as regards such current, is short-circuited by the junction.

As the accuracy of this method depends on the difference in temperature between the hot junction and the rest of the circuit, the hot junction should be protected from draughts. To increase the sensitivity the thermo-junction may be enclosed in an evacuated glass bulb. The minimum loss of heat by conduction and convection will then be experienced, and the greatest rise in temperature for a given heating effect will be obtained.

(*d*) A wavemeter circuit employing a two-electrode valve and galvanometer as indicating-device is shown in Fig. 6

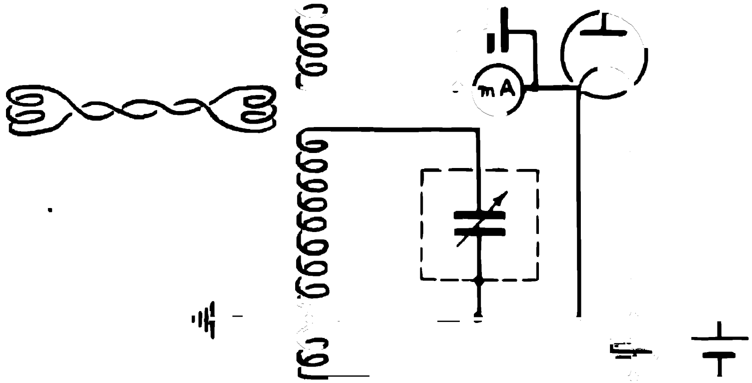


FIG. 6.

The diode acts in its usual manner as a rectifier, *i.e.*, current only flows when the anode is positive to the filament. The D.C. milliammeter in the external circuit from anode to filament reads the mean value of the anode current pulses. This will have its maximum value when the P.D. between anode and filament is a maximum, provided that saturation current is not reached.

The E.M.F. is introduced into the wavemeter circuit by an untuned intermediate coupling circuit. Maximum current flows in the wavemeter circuit when it is tuned to this E.M.F., and the maximum flux then produced induces maximum E.M.Fs. in the coils which give the anode voltage and filament current. When the filament current produced by this means is insufficient, a cell is provided which may be switched into the filament heating circuit.

(*e*) Fig. 7 represents a wavemeter circuit employing a triode and milliammeter. The wavemeter tuning is indicated by the deflection of a milliammeter in the anode circuit of a detector valve. The variable condenser is turned until a maximum deflection is obtained. The detector valve is used as an anode bend rectifier with sufficient negative bias to prevent grid current in normal use. The potentiometer across the H.T. supplies the grid bias for the valve, and by increasing this grid bias a larger " signal " is required between grid and filament to give a deflection on the milliammeter. This adjustment provides the coupling control in the instrument between the tuned circuit and the indicating device.

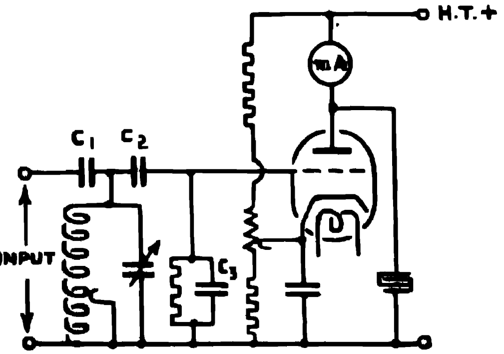


FIG. 7

If the wavemeter is severely over-loaded by an excessive signal, the large grid swung on the detector valve will cause a grid current. This grid current, passing through the grid leak, will increase the negative bias and move the working point further towards the " cut-off point," and so

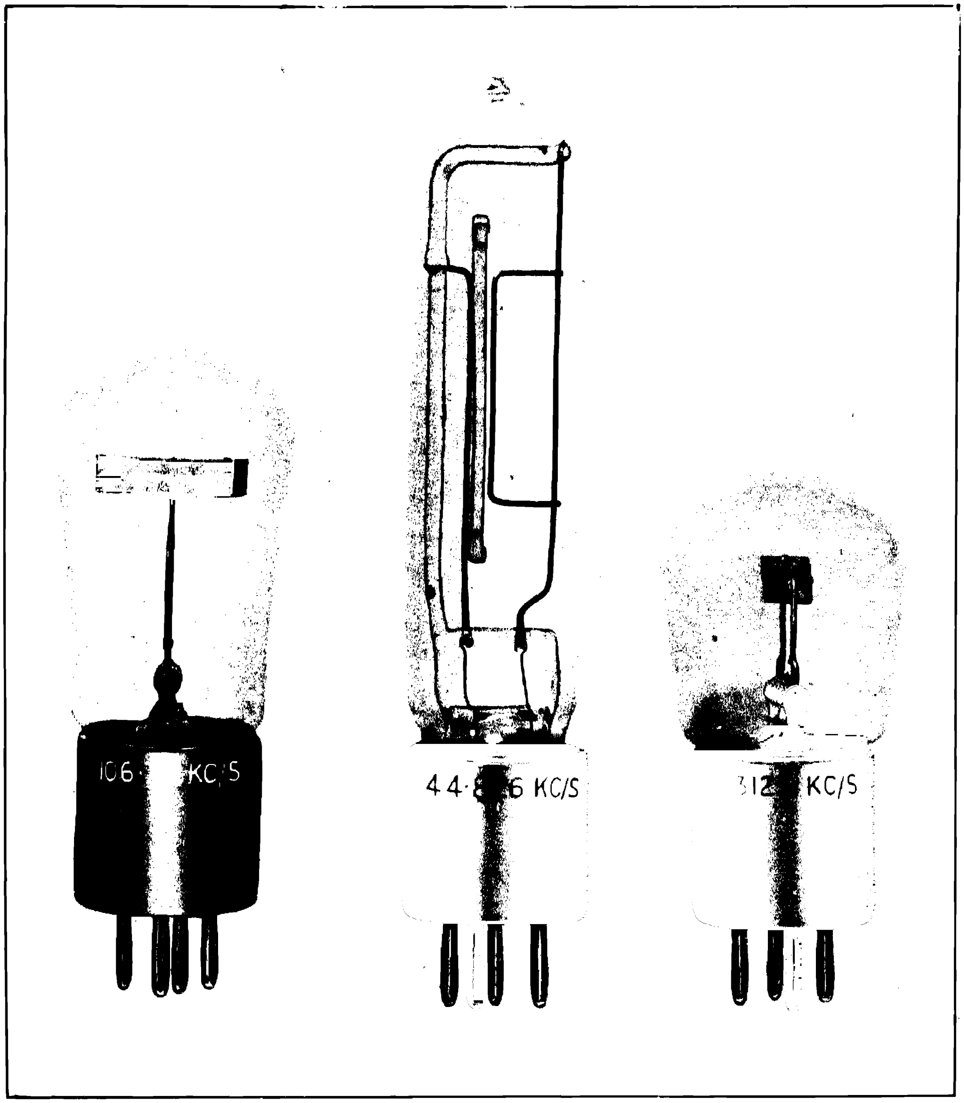


FIG. 8

reduce or limit the rectified anode current. This arrangement protects the milliammeter from a severe over-load which might burn it out. Very frequently, a special damped movement is fitted to the milliammeter to eliminate, as far as possible, the needle flicker due to variations of output from the transmitter caused by supply voltage fluctuations. Needle flicker makes it very difficult to tell when maximum deflection is reached.

(*f*) Luminous quartz resonators, invented by Giebe and Scheibe, have been in use in Germany for several years. These resonators are visual indicators of frequency and may be used to check the performance of calibrated circuits, or of a transmitter; they depend for their operation on the well-known piezo-electric properties of quartz crystals. A crystal is lightly held in a metal fitting which constitutes one of the electrodes in a gas discharge tube, the other electrode being a wire or small metal plate near the surface of the crystal, Fig. 8. The usual gas filling is neon at about 3 to 5 mm. pressure. The metal fitting, holding the quartz, is constructed so that no part of it is as close to the other electrode as the surface of the quartz.

With the electrodes connected appropriately to a resonant circuit, powerful oscillations will cause the neon to glow brightly with the usual insensitive glow discharge between the electrodes. If, however, the input is considerably reduced in strength, so that direct glow discharge between the electrodes cannot be produced, then a much fainter but very distinct glow appears on the surface of the quartz when its resonant frequency is applied. This characteristic glow is due to the field produced by the piezo-electric charge on the surface of the quartz, and the sensitivity of this visual indication of oscillations is about $\pm 0\cdot 02$ per cent.

It has been found convenient to mount the resonators in ordinary 4-pin valve bases, the electrodes carrying the quartz being connected, one to the grid pin and the other to the anode pin. Fig. 8 shows three samples. The short one, with a horizontally supported bar of quartz, is similar in general appearance to the original German models and is typical of resonators with frequencies of about 100 kc./s. The largest type was constructed for lower frequencies of 40-50 kc./s. The one with the square crystal slice is designed to work on a frequency of about 3,000 kc./s.

7. Comparison of Indicating Devices.

(*a*) Pea lamp. Capable of moderate accuracy; cheap and robust.
(*b*) Neon tube. Capable of moderate accuracy; fairly cheap and robust.
(*c*) Vacuum thermo-junction and galvanometer. Capable of high accuracy; expensive and fragile. Slow in action.
(*d*) Diode. Capable of high accuracy; cheap and robust.
(*e*) Triode and milliammeter. Capable of very high accuracy; cheap and robust. It is more sensitive than a diode and is much more commonly used. In the Service, it has completely replaced the diode.
(*f*) Luminous quartz resonator. Fairly cheap and robust; gives accurate indication of resonance to about one part in 5,000.

8. **Factors Affecting Wavemeter Accuracy.**—The certain accuracy with which a frequency can be measured with a wavemeter is determined by :—

(*a*) The frequency discrimination of the instrument.
(*b*) The constancy of the variable condenser and other components.
(*c*) The temperature coefficients of the various circuit components.
(*d*) The exterior coupling between the wavemeter and the transmitter, or receiver.

9. **Frequency Discrimination.**—This depends on a number of related factors, the most important of which are :—

(*a*) The precision with which the variable condenser dial can be read.
(*b*) The sharpness of tuning.

(*a*) Accuracy of reading depends on the relative size of the condenser scale, with reference to the capacity range which it has to cover. To assist in accurate reading, condensers are usually provided with slow-motion devices, and sometimes with verniers.

(*b*) Frequency discrimination depends very much on the accuracy with which the resonant frequency may be determined, and this depends on how sharply the current falls on either side of its peak value at resonance, *i.e.*, on the selectivity of the wavemeter circuit.

It is shown in Vol. I that selectivity is governed by two quantities, the ratio of inductance to capacity (L/C), and the total resistance or damping (R).

The question is summed up by saying that, for high selectivity, the " power factor " of the coil $\left(\text{approximately } \frac{R}{\omega L}\right)$, or " log decrement " $\left(\frac{\pi R}{\omega L}\right)$, should be small. Typical " power factors " in this work range from (say), 0·0015, for a small inductance, to 0·006 for one of the order of 1 henry.

The resistances in the circuit are the high frequency ohmic resistance of the inductance, connecting leads and indicating device, and the equivalent resistance of the power losses in the condenser. With high quality components, the principal cause of damping may sometimes be the indicating device. It may be reduced in various ways. In some cases it is done by arranging the indicating device in a separate circuit loosely coupled to the main wavemeter circuit, as shown in Fig. 3 ; the equivalent resistance of the lamp is then less than its actual resistance by a factor depending on the coupling. With a thermionic tuning indicator, such as a triode and milliammeter operated as an anode bend rectifier, the damping resistance in parallel with the tuned circuit is reduced to a negligible amount so long as the conditions of operation do not involve the flow of grid current. With an indicator as sensitive as this, it is possible to obtain the necessary sharpness of tuning, using small and robust inductance coils having a relatively high power factor, rather than by the use of large and delicate tuning coils. In one Service case, power factors range from 0·01 to 0·03.

The nature of the indicating device plays a large part in determining within what limits the frequency for maximum current may be estimated. A pea lamp or neon tube, for example, in which the brilliancy of the light is the criterion, does not lend itself to such accurate determination as an instrument in which a pointer reading may be observed. Moreover, in the case of a wavemeter employing a thermo-junction and galvanometer, a short time elapses between the heating effect of the high frequency current producing the temperature and the thermo-electric E.M.F. corresponding to it, and, accordingly, the pointer reading lags slightly on the wavemeter condenser variation. A thermionic tuning indicator is free from most objections ; it is very sensitive and has no inertia.

It is important in wavemeter work to distinguish between the two terms, **frequency discrimination** and **absolute accuracy.** This distinction may be made clearer by means of the following figures. The standard frequency measuring equipment used in H.M. Signal School has a frequency discrimination of the order of one part in 150,000 ; the absolute accuracy of the reference frequency given by means of a quartz crystal under thermostatic control, is of the order of a few parts (three or four) in 10^7.

A typical modern portable absorption wavemeter, in use in the Service, has a frequency discrimination of the order of one part in 2,000. It has an absolute accuracy, including temperature effects, of about one part in 300.

10. **The Constancy of the Variable Condenser and other Components.**—It is not, in general, necessary that the inductance and capacity of a wavemeter circuit should be accurately known, since the ordinary wavemeter is calibrated as a circuit by comparison with a standard. The question of how the standard is obtained will not be considered here. The important point is that the values of the components should not vary from those they possessed in the circumstances under which the calibration took place. The inductance and capacity of the indicating device, that between the leads, and between the circuit and its surroundings, thus become important in spite of their small values.

With fixed inductances, most of these effects can be kept constant as far as the total inductance of the circuit is concerned. When the operating E.M.F. is injected into the circuit by means of a mutual coil connected to the wavemeter circuit by flexible leads, the length of these should not be altered, nor should they be gathered in a coil for convenience. By keeping them extended at full length, their inductance should be the same under all conditions. In modern wavemeters, to prevent such alterations, the leads are made stiff, and with wavemeters forming an integral part of a modern transmitter, the necessary coupling is under precise control. For an accurate instrument, the use of a coupling coil with long flexible leads is most undesirable; it is only permissible in a second grade instrument employed in measuring low and intermediate frequencies.

The capacity presents a more difficult problem. Changes in the value of the variable condenser itself, owing to mechanical wear, may be obviated by good design and construction; a condenser may usually be assumed to maintain its accuracy to one part in 1,000 for a reasonable period, including temperature changes. The effect of stray capacities to earth of the various parts of the circuit is more difficult to deal with and generally constituted the final limitation on the accuracy of the older unscreened wavemeters.

Every point not at earth potential in the wavemeter circuit has capacity to earth, and this may be large enough to be cumulatively appreciable. The stray capacity of an unscreened wavemeter varies with the position of the wavemeter circuit with regard to its operator (the well-known phenomenon of "hand capacity"), other apparatus in the room, and the walls of the room itself.

Since stray capacity cannot be avoided, the principal object in wavemeter design is to keep its value constant and independent of the position of the wavemeter with respect to neighbouring objects. This may be achieved to a very large extent by suitable screening. The stray capacity is then, almost entirely, capacity to the earthed metal screen, usually shown in wavemeter diagrams by means of dotted lines.

The capacity of the indicating device must also be taken into account. For example, the neon tube in Fig. 4 is in parallel with the wavemeter condenser in use. This, or any other parallel capacity, would automatically be taken into account in the process of calibration, and would not matter if it remained constant. Its capacity, however, depends on the distribution of the electric field between electrodes, and this varies with the brilliancy of the glow. The glow at resonance should be adjusted to the same brilliancy as it had under the conditions of its calibration; this is generally the minimum glow that can be observed. The adjustment will be effected, of course, by varying the coupling between the source of oscillations and the wavemeter.

In the case of the triode valve and milliammeter, the tolerated differences of inter-electrode capacity, in valves of the same pattern, constitute an important source of error.

The effect on the frequency determination, of a change in instrumental capacity of this nature, is most marked when the minimum wavemeter condenser setting is being used; the uncertain capacity is then a larger proportion of the total capacity. If the wavemeter is one of simple type, with a neon tube indicator, the frequency determination should be made with that value of inductance which necessitates the largest possible condenser setting.

In the wavemeter circuit of Fig. 7, to reduce the effect of changing capacities in the detector valve a capacity potentiometer, C_2 and C_3, is used as coupling between the tuned circuit and the grid of the valve. The ratio of the capacity C_2 to capacity C_3 is carefully arranged so that the possible error, arising from a change in the valve capacity in parallel with C_3, is maintained within specified limits.

11. The Temperature Coefficients of the Various Circuit Components.—A change in temperature of the surroundings in which a wavemeter is operated can produce an error which is sometimes seriously large. The temperature coefficient of the instrument might be defined as "the error introduced into a frequency determination by a unit change of temperature." The temperature coefficient of the whole is determined by that of its component parts.

The temperature coefficient of inductance depends on the materials employed and the shape of the coils. With careful design, it is possible to produce inductances having temperature coefficients not exceeding one part in 1,000 between working temperatures of 15° C. and 50° C.

It can be shown by a mathematical treatment similar to that in paragraph 5, that an error of this nature should not cause an error in frequency greater than one part in 2,000.

In the case of a condenser, a change in the dimensions of the condenser with change of temperature produces a corresponding alteration in capacity. It has been shown that between working temperatures of 15° C. and 50° C., the temperature coefficient for a brass condenser would be about one part in 1,500, producing a corresponding frequency change of one part in 3,000. For an aluminium condenser, the temperature coefficients are bigger.

Warping of any metal parts, or leads, may alter the tuning capacity by an amount only to be determined by experiment. Wherever possible in wavemeter construction, coupling capacities such as C_1, C_2 and C_3 of Fig. 7 should be of the temperature coefficientless type.

The frequency change produced by (say), an increase in temperature, is in the same " sense " for the changes produced both in the inductance and the condenser. The sum of the two errors in frequency produced by changes in the capacity and inductance forming the tuned circuit does not, however, account for a quite considerable error in frequency which is observed experimentally when wavemeters are subject to change of temperature. In most cases this discrepancy may be traced to the variable condenser, and is almost certainly due to the warping of the plates. These discrepancies are not regular ones and no real coefficients can represent them.

Changes of calibration with a temperature of the order of one part in 250 are quite commonly observed for a 30° C. alteration of temperature ; in all cases it is necessary to apply a correction when observations are taken at temperatures seriously different from that at which the calibration was made.

Instructions for applying a correction for an error of this nature are usually given with the curves expressing the results of calibration—the calibration curves. With an ordinary inverted frequency scale, one in which the frequency increases with the scale reading, the effect of a rise of temperature is to displace the calibration curve to the right. The actual frequency corresponding to any scale setting will then be **smaller** than that given by the calibration curve by an amount depending on the change of temperature. Typical figures for a particular type of Service wavemeter are the following :—

A condenser setting corresponding to 100 kc./s. at 78° F. means 99·6 kc./s. at 126° F.

A condenser setting corresponding to 20,000 kc./s. at 78° F. means 19,900 kc./s. at 125° F. These figures correspond to overall temperature coefficients of about one part in 6,750 and one part in 5,400 per degree Centigrade respectively.

12. **The Exterior Coupling between the Wavemeter and the Transmitter or Receiver.**—It is important that the only E.M.F. acting in the circuit should be that which comes from the source. Unwanted " pick-up " from extraneous sources can only be avoided by the most careful screening of all connecting leads, as well as the instrument itself.

It is also very essential that the coupling between the wavemeter and the transmitter or receiver should be a loose one. If it is otherwise, the wavemeter and the circuit to which it is coupled, virtually act as one circuit, the resonant frequency is altered and there may even be a pronounced double frequency effect. Calibration curves of the wavemeter would no longer be valid. It is for this reason that small portable wavemeters using (say) pea lamps as indicators, should never be placed so close to an oscillating circuit that the pea lamp glows very brightly ; far greater accuracy of reading will be given with loose coupling, and the glow of the pea lamp barely visible.

In the case of wavemeters which form an integral part of transmitters, the requisite looseness of the coupling will be an essential feature of the design.

13. **A Modern Type of Service Absorption Wavemeter.**—The simplified circuit diagram of Fig. 9 represents the general electrical features of a portable absorption wavemeter used for tuning transmitting sets, or, in conjunction with a local oscillator, as a heterodyne wavemeter for measuring the frequency of incoming signals, and calibrating receiver outfits.

The circuit consists of a coupling wave (1), the output of which is loosely coupled to a tuned circuit (3) and (4) which is, in turn, loosely coupled to a detector valve (2). The grid of the coupling valve (1) is connected to the transmitter, or other source of oscillations, by connecting a lead between the transmitter and the input terminal (5) on the wavemeter. The wavemeter tuned circuit consists of a tapped inductance and a variable condenser (0·5 jar). It will be seen that the tuned circuit is linked both to the coupling valve and to the detector by means of the capacity potentiometers formed by condensers (6)/(7) and (8)/(9) respectively. The use of the latter, in the case of the

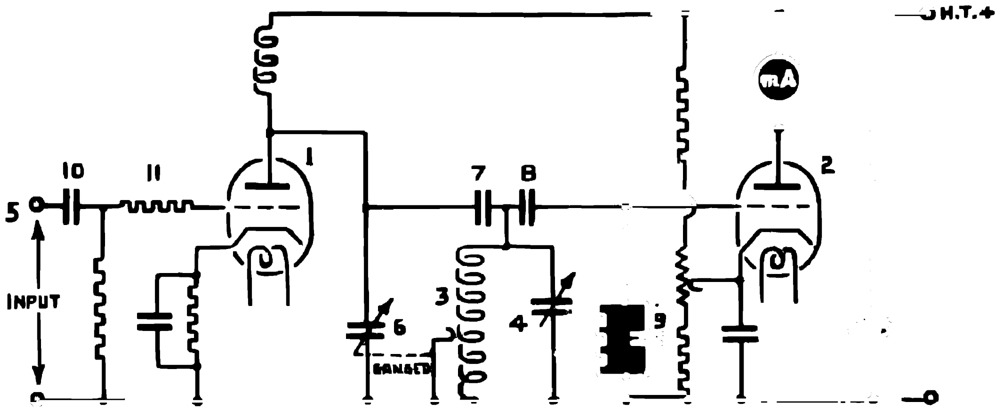


FIG. 9.

detector valve, has already been referred to in paragraph 10; the ratio of the capacity of (8) and (9) is important, since a minimum ratio must be maintained in order to prevent changes in the constants of the detector valve from affecting the calibration of the wavemeter. Similar considerations of accuracy impose a minimum ratio on the capacities of condensers (7) and (6). In the diagram, condenser (6) is shown as being variable; this enables the input to the tuned circuit to be adjusted as the frequency alters, and, in practice, condenser (6) consists of a number of fixed condensers. Condenser (6) is fixed for any one range; it varies from range to range and, for this reason, the diagram shows the "coupling condenser selector switch" ganged to a tap on the inductance. The representation is not a perfect one, and does not show that the same condenser sometimes does duty for two or more ranges. The condenser coupling prevents the complications which would arise if mutual coupling with coils were used, and the selection of suitable condensers (6) for each range position is equivalent to changing the coupling coils in a system utilising mutual coupling.

When the wavemeter tuned circuit is in resonance with an incoming signal, maximum oscillatory voltage will be developed across it and will produce maximum reading in the D.C. milliammeter in the anode circuit of the detector valve (2), in the way to which reference has already been made in paragraph 6.

The triode valve and milliammeter introduces so little damping into the tuned circuit that, in practice, it is found possible to design the latter employing robust inductance coils of relatively small diameter, and with a relatively high "power factor" (paragraph 9).

Coupling to the source of oscillations is made by a single lead from the coupling terminal (5) taken to some point at high oscillatory potential, to which it is coupled by a very small capacity. With a powerful transmitter, the stray capacity to a few inches of lead from the terminal (5) is sometimes sufficient; in very low power sets it can sometimes be obtained by twisting a few inches of insulated coupling lead around the aerial wire. Condenser (6) also constitutes the fine adjustment coupling control within the wavemeter. In order to limit a change in effective capacity of the wavemeter circuit caused by changes in the capacity to earth between the coupling valve and the coupling lead from terminal (5) to the transmitter, a fixed condenser (10) is introduced into the lead from terminal (5) to the grid of valve (1). Its capacity is usually of the order of 40 $\mu\mu f$. Resistance (11) is an "anti-parasitic" resistance, inserted to prevent spurious oscillations.

With the help of a local oscillator, the wavemeter employing this circuit may be used as a heterodyne wavemeter to measure the frequency of a distant incoming signal. The procedure is to tune the local oscillator, which may be used as a receiver, until its frequency is the same as that of the incoming signal; this is done by tuning the oscillator to the "dead space" (zero beat with the signal). Having tuned the oscillator to the same frequency as that of the signal, it is only necessary to couple the oscillator to the wavemeter input, in order to measure the frequency of the incoming signal. This, in principle, is the method involved, but in practice it is not always easy to tune the oscillator to a "dead space"; this is particularly so in the case of M/F, where the tuning is coarser.

With careful use, a wavemeter of this kind is capable of an accuracy of about one part in 500, including temperature change effects.

14. **Oscillators.**—With reference to paragraph 1, a local roughly calibrated oscillator is required for the following purposes:—

(*a*) In conjunction with a wavemeter, for setting a receiver to a desired frequency. The wavemeter is used to tune the oscillator that injects the signal by which the receiver is tuned.

(*b*) As an accessory for use with a wavemeter, to assist in the process of measuring the frequency of an incoming signal from a local or distant transmitter. This process is one which, in technical details, will differ with the nature of the oscillator.

Oscillators are merely roughly calibrated low power transmitters. Their circuits have the same distinguishing characteristics as those which serve as a basis of classification for transmitters (Section K).

15. **Oscillator with a Self-Quenching I.C.W. Circuit.**—Fig. 10 represents the electrical features of a C.W. or I.C.W. oscillator, the audio-frequency modulation, in the case of the latter, being produced by a "squegging" or self-quenching method. The self-oscillatory circuit is of

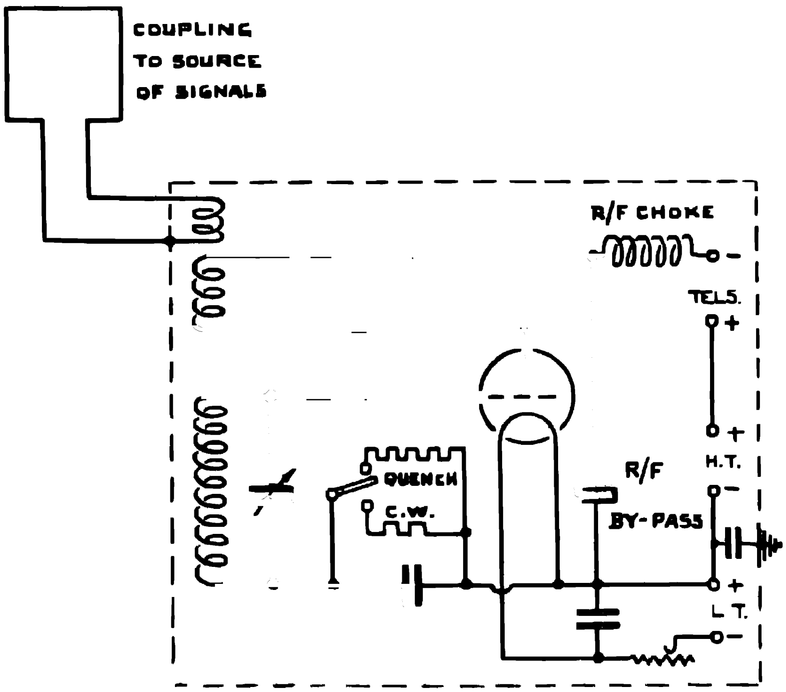
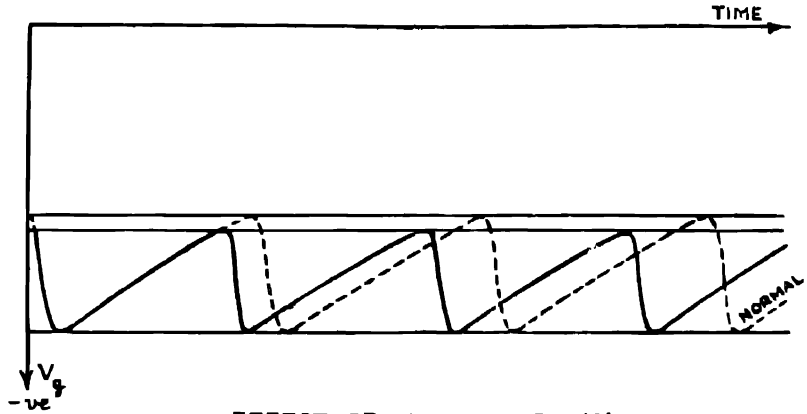


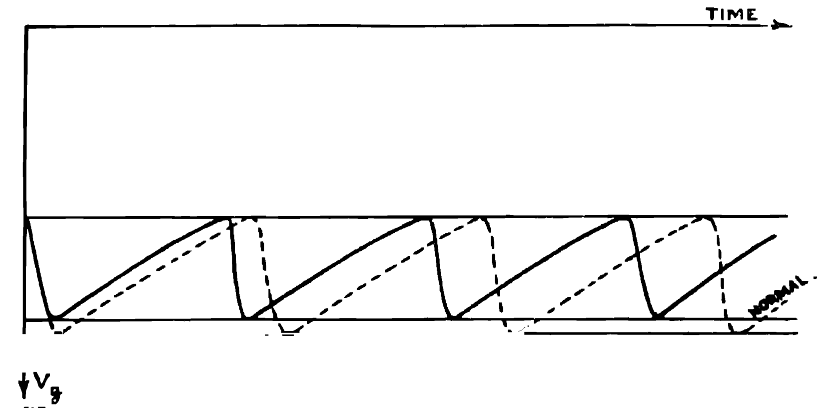
FIG. 10.

conventional type, the feed back to the tuned grid being obtained from the coupled coil in the anode lead. Two grid leaks are provided, either of which may be introduced by means of a 2-way change-over switch marked "quench," and "C.W." One grid leak has a much higher resistance than the other. When the leak of lower resistance is in circuit, C.W. oscillations are produced; with the leak of higher resistance, the grid runs so far negative that I.C.W. oscillations are produced with an audio-frequency modulation, the actual frequency of which depends upon the CR value of the grid leak and condenser combination.



EFFECT OF INCOMING SIGNAL

FIG. 11.



EFFECT OF INCREASE OF DAMPING

FIG. 12.

An instrument of this nature may be used as an oscillating receiver in both the "quench" and "C.W." positions. In the C.W. position, the oscillator acts as an autodyne receiver, *i.e.*, it is mis-tuned by an audio-frequency amount to the incoming signal. In the **quench** position, the presence of an incoming signal is indicated by a rise in the pitch of the note heard in the telephones, the rise in pitch being a maximum when the oscillator is tuned to the incoming signal.

The cause of the rise in the pitch of the note heard in the telephones which enables an oscillator in the quench position to detect incoming signals, is treated more fully in the section dealing with Super-regenerative Amplification (Section F). As a receiver, it is sometimes called a "howling squegger." Briefly, its action is due to a decrease in the audio-frequency period of grid voltage variation, produced because the energy supplied by the incoming signal allows the circuit to start self-oscillations at a more negative mean grid voltage than when all the energy must be suppiied by the oscillatory circuit itself. This is illustrated in Fig. 11.

The effect of "hurrying up" the audio-frequency cycle of operations would also be produced if the oscillations could be quenched at a more positive grid voltage than normal, provided that the mean grid voltage for the commencement of oscillations remains the same. There would then be a smaller negative charge on the grid when oscillations were quenched, and a shorter period would therefore elapse before this charge had leaked away sufficiently to bring the grid up to the potential at which oscillations would start to build up again. The audio-frequency period of grid voltage change would be decreased, and a rise in the pitch of the note heard in the telephones would be observed. This case is illustrated in Fig. 12.

This effect may be produced, in practice, by increasing the damping of the grid oscillatory circuit by coupling a tuned circuit to it. The extra resistance does not noticeably affect the grid voltage at which oscillations start. The condition for generation of oscillations is $\frac{Mg_m}{C} > R$. The slope of the static mutual characteristic, g_m, is increasing so rapidly at this grid voltage that an undetectably small decrease in negative grid voltage enables the condition to be fulfilled, in spite of the greater value of R.

The effect of the increased damping in altering the grid voltage at which self-quenching takes place is much more marked; g_m is, then, the average slope of the dynamic characteristic, and this varies very little with change in mean grid voltage at large negative grid voltages. It thus becomes too small, to enable oscillations to be maintained, at appreciably more positive mean grid potentials as the resistance of the oscillatory circuit increases. Thus, the greater the damping, the higher is the pitch of the note heard in the telephones.

Advantage is taken of this fact to calibrate an oscillator producing I.C.W. by self-quenching action.

If a wavemeter is magnetically coupled to the oscillator working in the quench position, maximum damping will be introduced into the oscillator circuit when the wavemeter circuit is in tune with it. Accordingly, adjustment is made to the variable condenser in the oscillator circuit, until the greater rise in pitch of the note heard in the telephones is observed. When this is the case, the two circuits are in tune at a frequency determined from the wavemeter calibration curves.

16. **Uses of an Oscillator and Wavemeter Combination.**—A wavemeter circuit designed for use both as an ordinary wavemeter and for working with the oscillator referred to above, is shown in Fig. 13.

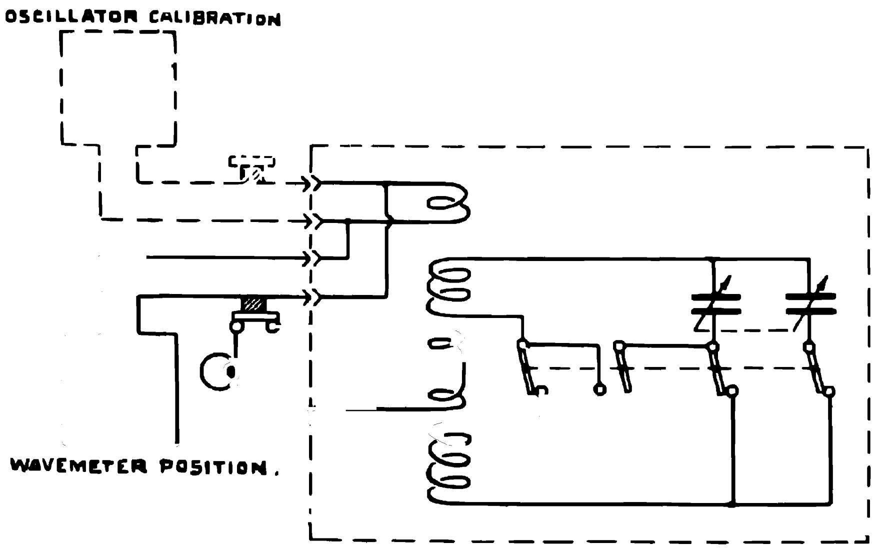


FIG. 13.

The coupling coil, with its stiff connecting leads, may be connected in either of two positions, as shown. The " full line " drawing shows the coupling coil in the " wavemeter position." A metal contact, which is carried on one lead but is insulated from it by an ebonite block, completes, in this position, the circuit containing the indicating device, a lamp. The switching arrangements in the main wavemeter circuit for altering the range should also be noted.

When used for calibrating the oscillator described above, the dotted line position of the coupling coil is used. The indicating device is, then, the telephones in the oscillator circuit, and the lamp is not required. The lamp circuit is therefore broken, to decrease the damping and so sharpen the tuning of the main wavemeter circuit. The wavemeter is set to a pre-determined frequency, and the frequency of the oscillator, in the self-quench position, is varied until the greatest rise in pitch of the note heard in the telephones is observed.

The two positions of the coupling coil necessitate two sets of calibration curves for this instrument since in one case the pea lamp circuit is closed and in the other it is open.

Once the oscillator has been calibrated in the above manner, it is a simple matter to tune a receiver to the same frequency. It is necessary to arrange a loose coupling between the receiver and the oscillator. The receiver is used in the oscillating condition and it is tuned until a note is heard. Using the telephones in the anode circuit of the oscillator, further adjustments are made to the tuned circuit of the receiver until a distinct rise in note is heard in the telephones.

The above method of tuning a receiver is inclined to be a little slow. A more rapid way, although less accurate, is to use the calibrated oscillator in the C.W. position. Using the phones on the receiver the signal from the oscillator is tuned in to the dead space. This procedure is less accurate than the one first described because the oscillator is calibrated in the quench position ; in changing to the C.W. position, a change of grid leak is involved and the error in frequency may amount to 1 per cent. If the receiver is of the non-oscillating type, the oscillator must be capable of producing I.C.W., *i.e.*, in this case it must be used in the quench position.

In the above process, the oscillating receiver constitutes, virtually, a low power transmitter. A transmitter, in the C.W. position, may thus be set to a pre-determined frequency by making adjustments on it until the telephones in the oscillator indicate the middle of the dead space ; the oscillator is used in the C.W. position.

A more accurate method is to use the oscillator in the quench position, and make adjustments on the transmitter to give a maximum rise of note in the " phones of the oscillator."

The process of measuring an unknown incoming frequency may be outlined as follows :—

(*a*) Tune an oscillating receiver accurately to the incoming signal. In the case of a C.W. transmitter, the receiver is adjusted until reaching the centre of the dead space. For an I.C.W. transmitter, the tuning is varied until loudest undistorted signals are heard in the telephones. In any other setting at least two notes will be heard, one, the I.C.W. note of the transmitter, and the other, a beat note between the transmitter and the local oscillator. Actually a confused " mush " of sound is heard. This " clears up " to give a true note when the oscillating receiver and the transmitter are in tune. This note is also the loudest, for when the transmitter and receiver are in tune, the maximum current will be produced in the tuned circuit by the E.M.F., due to the transmitter.

(*b*) With the oscillator in the quench position, adjust the frequency until a rise in the pitch of the note due to the oscillating receiver is observed in the telephones of the oscillator. Adjustment is continued until the rise in pitch has its maximum value.

(*c*) The oscillator is now coupled to its appropriate wavemeter, and the latter is adjusted until a maximum rise in pitch of note is again observed in the 'phones of the oscillator.

(*d*) From the setting of the wavemeter, and its calibration curves, the exact value of the frequency of the incoming signal may be determined.

When the incoming signals are strong, the frequency may be measured directly by using the oscillator as a receiver.

17. **Oscillator, with Separate A/F Modulating Valve.**—Fig. 14 represents the simplified circuit details of an oscillator, designed for work in conjunction with the wavemeter represented

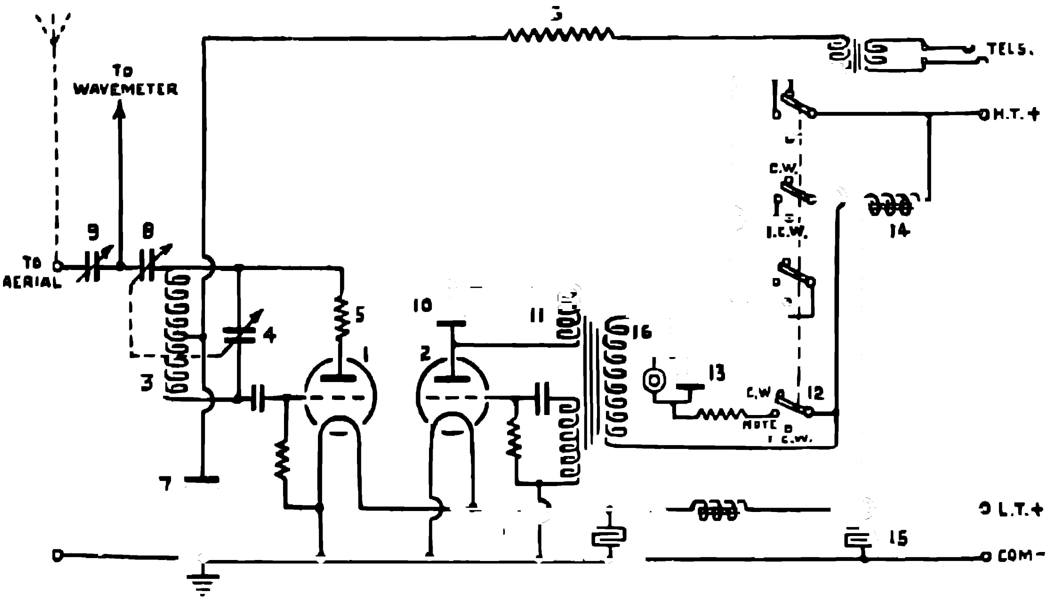


FIG. 14.

by Fig. 9. It differs from the oscillator described above, in the circuit arrangements by which self-oscillations are produced, in the method of producing I.C.W., and quite considerably in its method of use.

The oscillator had to cover a large frequency range, and, for this reason, a Hartley circuit was adopted. In standard terms, the circuit is described as having a tuned circuit (3) and (4) between anode and grid, direct inductive grid excitation ; it is also series fed. The grid tapping has been kept as low as possible in order to minimise the proportion of higher harmonics in the output. The small resistance (5) in the anode lead to valve (1) is there in order to prevent parasitic oscillations in the connecting lead. The resistance (6) and R/F by-pass condenser (7) together constitute the decoupling arrangements. The R/F oscillatory circuit is coupled to the wavemeter through a small condenser (8), and to the receiving aerial by a small condenser (9). Condenser (4) is ganged to condenser (8) in order to reduce the coupling to the wavemeter as the oscillatory circuit becomes stiffer.

Valve (2) is an audio-frequency oscillator, and coil (16) serves to modulate the H.T. supply to valve (1) when it is desired to produce I.C.W. The circuit is of conventional tuned anode type, with mutual inductive grid excitation. The anode coil (11) in parallel with condenser (10) gives a fixed tuning of approximately 500 cycles per second. Coupled to the anode and grid coils is an additional winding (16) on the iron cored coil, across which is connected a telephone earpiece tuned by condenser (13) to 500 cycles per second ; the functioning of this earpiece is controlled by a key, and the object of the fitting is to produce an audible 500 cycle note in air, to obviate the necessity for having a tuning fork to do the same thing. The H.T. supply to the audio-frequency valve (2) is decoupled by choke (14) and the large electrolytic condenser (15) ; the low tension supply is similarly decoupled. Telephones are provided in order that the oscillator may be used as a heterodyne receiver.

This instrument has the same uses as those of the oscillator already described, but its mode of operation is simpler. Since it is not calibrated, it must always be used in conjunction with its appropriate wavemeter.

To calibrate or set a receiver, the desired frequency is first set on the wavemeter, with the help of its calibration curves. The wavemeter is coupled to the oscillator and the latter is adjusted to give maximum deflection in the milliammeter; this sets the oscillator at the same frequency as the wavemeter. Finally, the oscillator is coupled to the receiving aerial, and the receiver is tuned to the incoming signal. Throughout this operation the oscillator should be in the I C.W. position, in order to produce an audible note in the telephones of the receiver without the necessity for heterodyne reception.

To measure the frequency of a strong incoming signal, such as that from a local transmitter, a simple method is to use the oscillator as a heterodyne receiver. Using telephones in the oscillator, the latter is adjusted to the dead space. The oscillator is then coupled to the wavemeter and the frequency measured in the usual way.

In general, the measurement of the frequency of an incoming signal involves the use of a separate receiver. The receiver is tuned to the signal frequency using its own heterodyne; in the case of I.C.W. it is tuned to the frequency of the carrier wave—paragraph 16 (*a*). The roughly calibrated oscillator is then substituted for the heterodyne belonging to the receiver, and, finally, the wavemeter is used to measure the frequency of the oscillator, *i.e.*, the frequency of the incoming signal. The process of adjusting the oscillator to the "dead space" is an easy operation at H/F but a more difficult one at M/F, where the tuning if relatively coarser. If the dead space is at all wide, it is difficult to determine the exact "zero beat" position in the middle of it and, in practice, a modification of the procedure is necessary for frequencies below about 100 kc./s. Instead of trying to tune the oscillator to the dead space, it is deliberately mis-tuned by an arbitrary amount in order to give an audible beat note. The oscillator frequency may be arranged to be either above or below the signal frequency; in each case, by adjustment the same note could be obtained. By taking the mean of the condenser settings giving the same beat note above and below the signal frequency, the exact position of the centre of the dead space can be obtained. Any convenient note could be taken, but in practice a 500 cycle note is used. To facilitate this adjustment, the telephone earpiece in the oscillator is set to emit a 500 cycle note in air; its function is only that of a tuning fork. If the beat note, as heard in telephones attached to the receiver, is out of tune with the 500 cycle note travelling in air from the earpiece, audible beats between the two will be produced. The operator then adjusts the frequency of the oscillator until the sensation of beats no longer is *felt*; with a little practice, this is an operation which is susceptible of the highest accuracy. It will usually be convenient to listen to the signal in the receiver using one earpiece only, leaving the other ear free for the reception of the note travelling in air from the "tuning fork equivalent" in the oscillator. The "note" is only required when the oscillator is in the C.W. position.

18. **Standard Frequency Transmissions.**—In order to facilitate the checking of the performance of calibrated apparatus, wavemeters, and some of the older directly calibrated oscillators used as heterodyne wavemeters, certain stations belonging to the three fighting Services transmit standard frequencies, to a definite programme, once or twice a month.

It is usual to set the transmitter as closely as possible to the desired wave frequency, and then to check the actual frequency of the transmitted signal by means of a standard wavemeter. A message is then sent out advising stations concerned what wave frequency has actually been transmitted.

The standard waves actually sent out are recorded, in the first instance, using a receiver, and the roughly calibrated oscillator for heterodyne purposes. This gives a series of "reference points," and further points may then be obtained as follows. Set any convenient C.W. transmitter accurately to zero beat with any one of the "spot frequencies" (say) 250 kc./s., obtained from the standard frequency transmission. Now set the oscillator to give approximately double this frequency. It will be found that a note can be heard in the 'phones of the oscillator which is due to the second

harmonic of the original oscillation ; all transmitters produce harmonics. A second reference point (500 kc./s.) can now be obtained by carefully adjusting the oscillator to zero beat with this second harmonic.

The new reference point corresponds to a wave frequency of twice the original frequency; others corresponding to three or four times the original frequency can be determined in the same way. Similarly, a further series of reference points may be obtained by tuning the transmitter to wave frequencies of 1/2 and 1/3, etc., of the known frequencies. In this manner the whole scale of wave frequencies used in the Service may be covered by starting from the few standard waves originally transmitted.

19. **Tuning of Emergency Spark Attachments.**—Most modern transmitters designed for use at sea, include a small emergency spark attachment. Its primary circuit may be tuned by means of a very loose coupling to any convenient wavemeter. Finally, the aerial circuit may be tuned by the simple process of adjustment until the aerial current is a maximum.

20. **Measurement of Aerial Inductance and Capacity.**—In a transmitter with a directly coupled aerial, the aerial constants form an integral part of the tuned LC circuit. Their value may be measured by making a known change in the value of either the inductance (L) or capacity (C), and measuring by wavemeter the resulting change in frequency. The method is as follows. The frequency of the aerial circuit is determined. The known alteration is then made in the aerial inductance, and a second frequency is determined. From these two results the two unknown quantities can be calculated as shown in the following example.

In a valve transmitter with 180 mics. in the aerial circuit, the resonant frequency was 600 kc./s., and with 300 mics. added it fell to 400 kc./s. Find the aerial inductance and capacity.

$$(L_{ae} + 180)\,\sigma = \frac{9 \times 10^8}{4\pi^2 \times 36 \times 10^4} = \frac{10^4}{16\pi^2}$$

$$(L_{ae} + 180 + 300)\,\sigma = \frac{9 \times 10^8}{4\pi^2 \times 16 \times 10^4} = \frac{9 \times 10^4}{4 \times 16\pi^2}$$

$$\therefore \quad 300\sigma = \frac{9 \times 10^4}{4 \times 16\pi^2} - \frac{10^4}{16\pi^2} = \left(\frac{9}{4} - 1\right) \times \frac{10^4}{16\pi^2}$$

$$= \frac{5 \times 10^4}{64\pi^2}$$

$$\therefore \quad \sigma = \frac{5 \times 10^2}{3 \times 64\pi^2} = 0{\cdot}264 \text{ jar.}$$

$$L_{ae} + 180 = \frac{10^4}{16\pi^2\sigma} = \frac{10^4 \times 3 \times 64\pi^2}{5 \times 10^2 \times 16\pi^2} = 240 \text{ mics.}$$

$$\therefore \quad L_{ae} = 60 \text{ mics.}$$

In the laboratory, the self-inductance and self-capacity of a coil may be measured by a method employing the same principle.

EXAMINATION QUESTIONS ON WAVEMETERS.

1. Discuss, in general terms, the factors influencing the overall accuracy of wavemeters. Describe the action of a thermo-junction and galvanometer as an "indicating device."
(Qualifying for Lt. (S), 1934.)

Section "W."

2. Describe a method of determining the self-capacity of a coil using a wavemeter. From the following measurements, taken in an experiment, deduce the self capacity of the coil :—

3 (in $\mu\mu F$)	10	20	30	40	50
λ (in metres)	3·67	4·49	5·16	5·76	6·32 $\times 10^2$

(Qualifying for Warrant Tels., 1934.)

3. Describe the construction and principle of action of a resonant type of wavemeter suitable for measuring the wavelength of a spark transmitter. What precautions should be taken to minimise error when using the instrument ?
(C. & G., Inter., 1933.)

4. Describe the construction of a simple type of wavemeter suitable for measuring the wavelength of a spark or I.C.W. transmitter. How would you use such a wavemeter in conjunction with a buzzer and inductance to compare the capacitances of two condensers ?
(C. & G., Preliminary, 1935.)

5. A wavemeter consists of a variable condenser having a range from 50 to 1,000 $\mu\mu F$, and two coils of 300 and 100 mics. respectively. If the coils are fixed so that their mutual inductance is 25 mics., what range will the wavemeter have when the coils are used in (*a*) series aiding (*b*) in series opposing ; (*c*) in parallel aiding ; (*d*) in parallel opposing ?
(I.E.E., 1933.)

THE DECIBEL AND THE NEPER.

1. **Historical.**—The "decibel" is the 1/10th part of a "bel" (after Alexander Graham Bell, inventor of the telephone sounder), a unit in which one may express power ratios, and gain or loss ratios of related quantities such as current and voltage. It originated in line telephony in 1923, when the American Telephone and Telegraph Company introduced a new unit, then called a "transmission unit"; this was to replace an older conception based on a ratio comparison between the decrease in signal strength produced by the given telephone line, and that produced by a "mile of standard cable" In 1924, an international advisory committee on long distant telephony in Europe, together with the representatives of the Bell system, agreed to recommend their countries to adopt as standards

EITHER the "**bel,**" a unit based on logarithms to the base 10, and equal to 10 of the American Company's "transmission units,"

OR the "**neper**" (after Napier), a unit based on Naperian logarithms to the base e.

The growth in popularity of the decibel, since 1929, has been so great that it is now almost a household word throughout all branches of Electrical Engineering and Acoustics.

2. **Definitions.**

THE DECIBEL:—Two powers P_1 and P_2 are said to differ by N "bels" when—

$$\frac{P_1}{P_2} = 10^N \text{ i.e. } N = \log_{10} \frac{P_1}{P_2} \quad \ldots\ldots\ldots\ldots \quad \text{in bels.}$$

Or in words—"The logarithm to the base 10 of the ratio of the powers, gives the gain or loss in bels." If $P_1 = P_2$, then $N = 0$.

In practice, a unit of one bel is found to be inconveniently large, and the 1/10th part of it—the decibel—is more often used. Using the smaller unit we have—

$$\frac{P_1}{P_2} = 10^{0 \cdot 1N} \text{ or } N = 10 \log_{10} \frac{P_1}{P_2} \quad \ldots\ldots\ldots\ldots \quad \text{in decibels.}$$

The basic power ratio is $10^{0 \cdot 1}$, that is, $1 \cdot 259$.

THE NEPER:—Two powers P_1 and P_2 are said to differ by N "nepers" when—

$$\frac{P_1}{P_2} = (e^2)^N \doteqdot 10^N$$

$$\therefore \left(\frac{P_1}{P_2}\right) = e^N \text{ i.e. } N = \tfrac{1}{2} \log_e \frac{P_1}{P_2} \quad \ldots\ldots\ldots\ldots \quad \text{in nepers.}$$

Any ratio in db. may be readily converted to nepers, for example 60 db.—

We have $\log_{10} \frac{P_1}{P_2} = 6$ bels.

$$\therefore \tfrac{1}{2} \log_e \frac{P_1}{P_2} = 2 \cdot 3026 \times 6 \times \tfrac{1}{2} = 6 \cdot 907 \text{ nepers.}$$

The neper is used in some European Countries, but is less commonly encountered than the decibel.

3. **Cables, Amplifiers and Attenuators.**—With these units, if the signal strength of a cable signal is 1/10th of that at the transmitting end, the loss is 1 bel. With two similar cables in series, the received signal would be 1/100th of the transmitted one, and the loss would be two bels.

In any amplifier, if the output power is 100 times the input, the "gain" is two bels or 20 db; with two such amplifiers used in series, the gain in power ratio would be 10,000 : 1, or 40 dbs.

One of the many advantages of the decibel is that the enormous ranges of power involved in communication work can be expressed in figures conveniently small, instead of astronomically big ones.

In the above example, it will be noted that the nett power ratio of two amplifiers in series involves the *product* of their individual power ratios, but only the *sum* of their decibel equivalents. This is due to the logarithmic nature of the unit employed.

An attenuator may be regarded as the converse of an amplifier, and its power loss may be described in similar units. If two stages of amplification, one of p and the other of q dbs. is followed by a line having attenuation of s dbs., the nett level at the output of the system will be (p + q — s) dbs. above the original input level.

4. **Voltage Gain in Dbs.**—In general, two powers P_1 and P_2 will be compared by observing, either, the voltage developed across a given impedance, or, the current through it. **If the input and output impedance of (say) an amplifier are equal,** the power ratio will be proportional to the square of the voltage (or current) ratio.

$$\therefore \quad \frac{P_1}{P_2} = \left(\frac{V_1}{V_2}\right)^2 = \left(\frac{I_1}{I_2}\right)^2$$

i.e. $$N = 10 \log_{10} \left(\frac{V_1}{V_2}\right)^2 = 20 \log_{10} \frac{V_1}{V_2} \ldots\ldots\ldots \text{ in dbs.}$$

$$\text{or } N = 20 \log_{10} \frac{I_1}{I_2} \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \text{ in dbs.}$$

The voltage or current ratio relation is very frequently misused to describe a power ratio, without regard to the necessary condition of equality between the input and output impedances. For example, it is not necessarily correct to say that an amplifier with a V.A.F. of 100, has a power gain of 40 db., when its "voltage gain" is given by that figure. Under proper conditions, however, an amplifier having a V.A.F. of 1000, or a power ratio of 1,000,000, would have a power (or voltage) gain of 60 db.

POWER AND VOLTAGE GAIN RATIOS EXPRESSED IN DECIBELS.

Dbs.	Power ratio.	Voltage ratio.	Dbs.	Power ratio.	Voltage ratio.
0·1	1·023	1·012	6·0	3·98	1·995
0·2	1·047	1·030	7·0	5·01	2·238
0·3	1·072	1·035	8·0	6·31	2·456
0·4	1·096	1·047	9·0	7·94	2·663
0·5	1·122	1·059	10·0	10·0	3·162
0·6	1·148	1·077	20·0	100	10·0
0·7	1·175	1·084	30·0	1,000	31·62
0·8	1·202	1·096	40·0	10,000	100
0·9	1·230	1·109	50·0	100,000	316·2
1·0	1·259	1·122	60·0	1,000,000	1,000
2·0	1·585	1·259	70·0	10×10^6	$3·162 \times 10^3$
3·0	1·995	1·412	80·0	100×10^6	10^4
4·0	2·510	1·585	90·0	$1,000 \times 10^6$	$31·62 \,..\, 10^3$
5·0	3·160	1·778	100·0	$10,000 \times 10^6$	10^5

Interpolation is easy; for example, 15·2db. = 10 + 5 + 0·2, which gives a power ratio 10 × 3·16 × 1·047 = 33·2.

5. **Absolute Power Above an Arbitrary Datum—Decibel Meters.**—It is clear that the decibel is only a relative unit of power level; it is a ratio. If it is possible to arrive at some standard power as a "zero level," or datum of reference, then the absolute power may be expressed in decibels

above the chosen datum. Various datum levels have been chosen for different purposes, and 1 milli-watt, 6 milli-watts, and 12·5 milli-watts have been used in many cases. The zero output level of 1 milli-watt is frequently chosen for telegraphy work in radio engineering. In some recent sensitivity tests on a Service receiver, an output of 1 milli-watt at the telephone terminals, when using two stages of note magnification, was taken to correspond to a good R9 signal at the detector valve. In the same tests, input voltages between aerial and earth terminals were expressed in db. above 1 micro-volt as datum (ref. R.30).

A receiver output power meter is essentially an A.C. voltmeter, calibrated in decibels with reference to the power passing into a load of definite impedance, contained within the instrument and replacing the normal one. With an impedance of (say) 1000 ohms and a zero level of 1 milli-watt, the voltage to be measured would be equivalent to

$$\sqrt{0 \cdot 001 \times 1000} = 1 \text{ volt R.M.S.}$$

For modern broadcasting telephony receivers, zero level in most output meters is fixed at 50 milli-watts. This is held to represent the weakest output to have any loudspeaker programme value. Scales are usually marked in milli-watts and decibels; powers less than 50 milli-watts would be given negative readings in dbs. For a maximum scale reading of 4000 milli-watts the power ratio is 80, and the point would therefore be marked 19·0 db. In good instruments the impedance may be adjusted to match the valve.

6. **The Human Ear Justifies the Decibel.**—The adoption of a new and unfamiliar unit to describe power ratios was not simply due to a desire to substitute small numbers for big ones; the origin of its size and logarithmic nature may both be traced to the peculiarities of the human ear. The power used by electrical apparatus employed in wireless, or other signalling services, is closely related to the power in the form of sound which it usually produces. From this inter-dependence between the twin sciences of acoustics and electrical signalling, it may readily be understood that the units employed by the one should be adjusted partially to fit the requirements of the other. The decibel is a particularly convenient unit for measuring ratios of sound intensities.

7. **The Unit Power Ratio and the M.S.C.**—Prior to 1923, in telephony practice the M.S.C. (mile of standard cable) was defined as " the difference in loudness in an 800 cycle note, perceived by an observer comparing the notes from two equal telephones, one at the input and the other at the output end of a mile of standard telephone cable." The observed difference in loudness corresponds very nearly to a difference in power of 20 per cent. and, from this, the suggestion arose to use a standard power ratio step of $10^{0 \cdot 1}$; this gives a difference of $10^{0 \cdot 1} - 1$, that is, about 25·9 per cent. This percentage change in the loudness of a monotone is an easily audible one; it is, however, only slightly greater than the minimum audible change of intensity. It needs a well-trained ear to detect a change in loudness of as little as 10 per cent., which is usually regarded as the lowest figure. Indeed, for most acoustic purposes, the decibel is defined as the least ratio of sound intensities that can be appreciated by the human ear.

8. **Logarithmic Response of the Human Ear—Fidelity Curves.**—An absolute measure of the intensity of sound of a pure tone travelling in air or other medium, can be obtained in various ways and expressed in terms of

(*a*) the R.M.S. pressure in dynes/cm.^{2} produced on a diaphragm placed in the path of the waves, or
(*b*) the amplitude of the oscillation in cms., or
(*c*) the number of watts or milli-watts crossing an area of 1 cm.^{2} at right angles to the direction of propagation.

If the loudness of a tone is increased, the amplitude increases, and also, the rate of dissipation of energy in the form of sound.

It is found that, no matter what may be the absolute value of the intensity of any sound, the sensitivity of the ear is such that the same percentage change of the original intensity will always produce the same relative alteration of the " loudness sensation " at the ear.

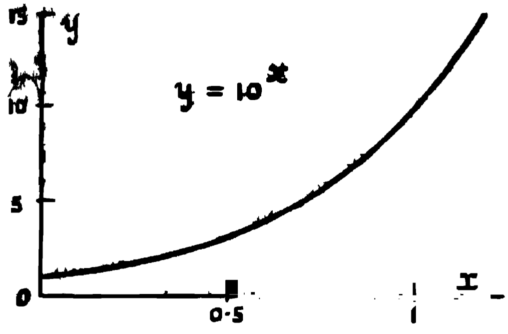
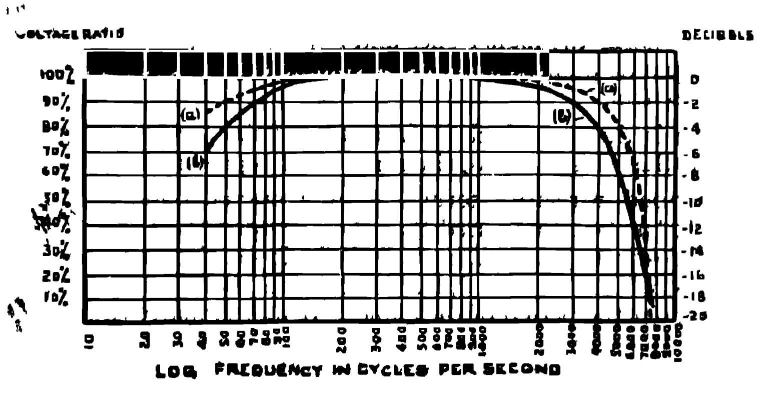


FIG. 1.

If an increase of intensity by 10 times produces a given effect at the ear, a further increase of 10 times, that is to say a total of 100 times, will produce twice the effect at the ear. Mathematically, this implies that the ear responds logarithmically to sounds of different intensities (*i.e.*, powers), or that a logarithmic graph will be obtained if " sound intensity " is plotted along the -y-axis, and " sensation of loudness " plotted along the -x-axis. The curve is similar to that obtained on plotting $y = 10^x$, Fig. 1. In a graph of this nature, equal percentage increases in the value of y give equal increments along the -x-axis. Considering the output of wireless receivers, the justification of the decibel notation is that a change in output from 8 to 40 milli-watts (7 db) would seem to the ear identical with the change from 40 to 200 milli-watts, since the power ratio is again 5 : 1.

It is interesting to note that, in addition to the aural response to power being logarithmic, the ear also responds logarithmically with respect to pitch (*i.e.*, frequency). The graph of frequency plotted against " musical interval " along the -x-axis would be similar to that of Fig. 1. Tones separated by an octave have a frequency ratio of 2 and are detected by the ear as being similar musical intervals The various C's on a piano would appear equally spaced along the -x-axis of a graph like that of Fig 1.

Fig. 2 shows an amplifier fidelity curve. It represents the differential amplification of all frequencies in the usual A/F range. Ideally, an amplifier should amplify all frequencies equally well, and, in that case, the fidelity curve would be a straight line. This curve shows that with reference to the output at 500 cycles, frequencies below 200 cycles are not equally amplified. The graph gives a true representation of the fidelity with which the amplifier treats the various frequencies. At one time, these graphs were plotted using a voltage ratio scale along the y-axis ; since the ear is logarithmic in action, curves of that nature cannot represent truly the aural effect, and give a wrong impression of the performance of the apparatus. For example, at 4000 cycles, the full curve shows the voltage ratio as being 80 per cent., or 1·25 taking the reciprocal ; corresponding to this voltage ratio, from the table we obtain approximately 2 dbs., and the dotted curve accordingly passes through the point -2 db.



FREQUENCY-RESPONSE CURVES OF AN AMPLIFIER

(a) IN DECIBELS
(b) IN VOLTAGE RATIOS

FIG. 2.

9. **Power in Sounds—Decibel Level—Tone Control.**—The sensitivity of the ear varies with the frequency and also with the " level " at which the sound is produced. For any normal person there is a minimum sound intensity for each frequency, below which nothing is heard. Fig. 3 is due to Fletcher and Munson ; the lower curve shows the relative variation in level of the

"threshold of audibility" over the ordinary A/F band. The level at 1000 cycles is arbitrarily marked 0 and is taken as a zero or datum level.

Ordinary conversation is approximately 60 db., above the threshold of audibility. The curves may be termed "equal loudness curves," and show the relative insensitivity of the ear to low notes when the intensity level is low. In the case of the curve 0, a just audible tone at 100 cycles has a sound power about 38 db. above a similar one at 1000 cycles.

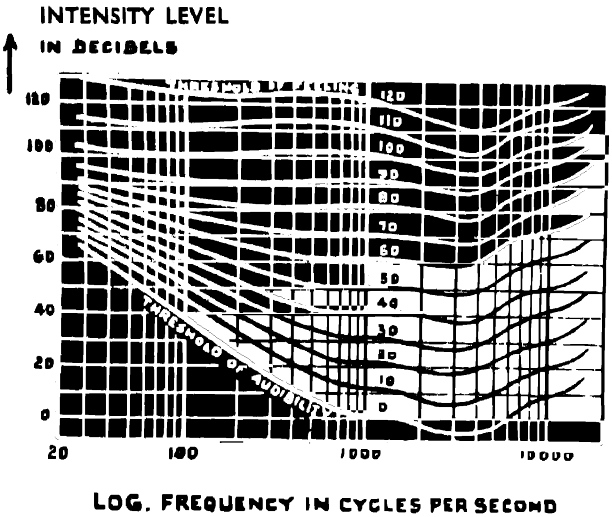


FIG. 3.

At an intensity level of 60 db. at 1000 cycles, an equally loud tone at 100 cycles would only differ in power by about 12 db. For very intense sounds over about 90 db., the ear appears almost uniformly sensitive over the above frequency range. Still more intense sounds are felt rather than heard, and there is, in fact, a boundary called the "threshold of feeling." From the practical point of view it follows, that a voice cannot sound natural unless it is reproduced at its natural level. When this is not possible, or desirable, the volume control is turned up and the intensity level is raised; although the relative power content may remain the same, the increased sensitivity of the ear to the lower notes produces the well-known and unpleasant sensation of "booming." In a large auditorium, a surfeit of low tones masks the higher ones and impairs the general intelligibility of speech. For this reason, in large public address systems, a volume control should always be operated in conjunction with a "tone control."

It will be noted that the 50 db. curve is the one which is flattest over the greatest range of frequencies, and it is sometimes considered that this is the optimum level of reproduction.

Sound powers have, so far, been described in relative terms. It is, however, possible to put each sound power on an absolute basis. It has been measured that the average power of ordinary speech is somewhere between 10 and 15 micro-watts; on this basis G.W.C. Kaye calculated that the continuous talk of a Wembley Stadium football crowd of 100,000 persons represented only enough energy to light a small electric lamp throughout the period of the game. It was similarly calculated that the acoustic disturbance created by a ship's syren amounts to about 6 micro-watts per square centimetre at a distance of about 115 ft. This represented a total dissipation of energy by sound of about 1/3rd of a horse power. The peak power of the loudest sound in conversation has been stated to be of the order of 5000 micro-watts, the power of the faintest sound being in the region of 0·01 micro-watts; this represents a range of about 57 db. The power corresponding to the datum line of audibility itself has been variously estimated by different observers. Although there is no general agreement about the exact figure, the number 10^{-16} watts per square centimetre, for a free progressive wave of frequency 1000 cycles per second, is now usually quoted by American Research workers as the datum line. With this level as a basis, the level of ordinary conversation appears as about 100 db. (not 60 db. as quoted above), assuming the average power of ordinary speech to be 10 micro-watts.

Musical sounds have a much greater power range than speech sounds. The peak power produced by a large orchestra may be of the order of 100 watts, and the noisiest drum can produce peaks of power of 25 watts.

Noises may, similarly, be expressed in terms of decibels above a datum. Measurements have been made over a complete scale of noise levels, from that of an aeroplane engine (110 db.) including

in decreasing order of loudness, a roaring lion (90 db.), an average office (45 db.), the purring of a cat (about 15 db.), and the noise in an underground vault in N.Y. city (10 to 15 db.).

Man-made noise and static of various kinds is usually present as a background during the reception of broadcast telephony signals. In order that any transmission should provide good programme value for the ordinary listener, a committee of the I.E.E. recently decided that the signal/noise should be not less than 40 db.

10. **The Loudness Unit. The British Standard Phon.**—The decibel serves as a useful unit for the expression of power ratios, but for the measurement of the relative loudness of sounds and noises, a new but related unit, the " phon," was introduced by the British Standards Institute in 1936.

The disadvantage of the decibel as a loudness unit is demonstrated by Fig. 3, a different decibel level being required to express equality of loudness at different frequencies in the A/F range. A unit is needed with which to describe relative loudness, and to express equality of loudness along each of the equal loudness contours. It was decided to express loudness in terms of the equivalent loudness of a standard reference tone, and to define a datum of equivalent loudness based upon the R.M.S. pressure at the threshold of audibility of a free progressive tone at 1,000 cycles per second ; that R.M.S. pressure has been standardised in this country at 0·0002 dynes per square centimetre. More accurately, it is an intensity level of 10^{-16} watts per square centimetre, corresponding to a sound pressure of 0·000204 dynes per square centimetre, at 76 centimetres of mercury and 20° Centigrade.

Loudness may be expressed as a ratio above this arbitrary datum, **the loudness of a sound in " phons " being numerically equal to the sound intensity in decibels of an equally loud 1,000 cycles per second pure note.** The equal loudness contours of Fig. 3 could be given labels in phons equal numerically to the decibel levels of each at 1,000 cycles per second.

In general, the loudness increases regularly with the intensity of the sound, within the limits of audibility, but, unfortunately, loudness is a *sensation*, the valuation of which depends upon *subjective* considerations such as the way in which the sound is heard, the person making the observations, etc. For example, in practice it is extremely difficult to say exactly when two sounds, such as that of a given continuous noise and a multiple of the standard reference tone, are equally loud ; in general, it is found relatively easier to decide when one sound is very slightly louder than the other.

A hypothetical " normal " observer can make an accurate decision as to equivalent loudness if " he " listens with both ears to the source of sound and the reference tone, both presented alternately from a position directly in front of the observer. A satisfactory approximation to this hypothetical being is achieved by taking the average of the subjective decisions of a group of 10 observers ; it has been estimated that such a group can arrive at a decision with an accuracy of $\pm$ 2 phons.

Subjective measurements of the above kind can usually only be made in a laboratory, and, for noise research work in engineering, it is very desirable to eliminate the personal equation as far as possible, by measuring noise by purely instrumental means. **Noise meters** usually consist of a microphone, amplifier, and indicating instrument ; they can be supplied with circuits enabling them to approximate in function to the human ear, and to give direct **objective measurements** of loudness which are accurate when applied to pure tones.

These meters may also be applied to the measurement of the loudness of continuous noises which vary in nature from pure tones to " multi equal loudness " tones. With complex noises, the reading of the meter may be seriously in error, and satisfactory relative measurement of noise can only be achieved if it is previously calibrated in a laboratory, by some subjective equality method ; with reproducible noises this is usually possible.

Objective noise meters are particularly valuable in dealing with noises of short duration, when there would not be time for any subjective measurement. Meters may be adapted for this purpose by the incorporation of suitable time constant circuits which conserve the energy produced during the sound impulse, permitting the meter to record a peak value from which it slowly falls.

In terms of phons the threshold of feeling has been estimated at 130 ; the noise level in motor-cars of various types has been estimated to range between 70 and 90 phons.

APPENDIX "B."

TABLE I.

The following conversion table is based on the formulæ :—

$$\lambda = \frac{3 \times 10^5}{f}, \text{ and } LC = \left(\frac{3 \times 10^4}{2\pi f}\right)^2, \text{ where}$$

f = frequency in kilocycles per second.
λ = wave-length in metres.
LC = oscillation constant, in microhenries and jars.

To obtain the LC value in microhenries and microfarads, the values given in the LC column should be divided by 900.

It should be noted that wave-length and frequency are reciprocal ; *i.e.*, 50 metres correspond to 6,000 kc/s., and 50 kc/s. to 6,000 metres.

Frequency (f) in kilocycles per second.	Wave-length (λ) in metres.	LC value (mic.-jars).	Frequency (f) in kilocycles per second.	Wave-length (λ) in metres.	LC value (mic.-jars).
300,000	1	0·0002533	6,977	43	0·4684
150,000	2	0·001013	6,818	44	0·4904
100,000	3	0·002280	6,667	45	0·5129
75,000	4	0·004053	6,522	46	0·5360
60,000	5	0·006333	6,383	47	0·5595
50,000	6	0·009119	6,250	48	0·5836
42,855	7	0·01241	6,122	49	0·6082
37,500	8	0·01621	6,000	50	0·6333
33,333	9	0·02052	5,454	55	0·7662
30,000	10	0·02533	5,000	60	0·9119
27,273	11	0·03065	4,615	65	1·070
25,000	12	0·03647	4,286	70	1·241
23,077	13	0·04281	4,000	75	1·425
21,429	14	0·04965	3,750	80	1·625
20,000	15	0·05699	3,529	85	1·830
18,750	16	0·06485	3,333	90	2·052
17,647	17	0·07320	3,158	95	2·286
16,666	18	0·08207	3,000	100	2·533
15,789	19	0·09144	2,857	105	2·793
15,000	20	0·1013	2,727	110	3·065
14,285	21	0·1117	2,609	115	3·350
13,636	22	0·1226	2,500	120	3·648
13,043	23	0·1340	2,308	130	4·281
12,500	24	0·1459	2,143	140	4·965
12,000	25	0·1583	2,000	150	5·699
11,538	26	0·1712	1,875	160	6·485
11,111	27	0·1847	1,765	170	7·320
10,714	28	0·1986	1,667	180	8·207
10,345	29	0·2130	1,579	190	9·144
10,000	30	0·2280	1,500	200	10·13
9,677	31	0·2434	1,429	210	11·17
9,375	32	0·2594	1,364	220	12·26
9,091	33	0·2758	1,304	230	13·40
8,823	34	0·2928	1,250	240	14·59
8,571	35	0·3103	1,200	250	15·83
8,333	36	0·3283	1,154	260	17·12
8,108	37	0·3468	1,111	270	18·47
7,895	38	0·3658	1,071	280	19·86
7,692	39	0·3853	1,034	290	21·30
7,500	40	0·4053	1,000	300	22·80
7,317	41	0·4258	967·7	310	24·34
7,143	42	0·4469	937·5	320	25·94

Appendix "B."

Table I—*continued*.

Frequency (*f*) in kilocycles per second.	Wave-length (λ) in metres.	LC value (mic.-jars).	Frequency (*f*) in kilocycles per second.	Wave-length (λ) in metres.	LC value (mic.-jars).
909·1	330	27·58	166·7	1,800	820·7
882·3	340	29·28	162·2	1,850	866·9
857·1	350	31·03	157·9	1,900	914·4
833·3	360	32·83	153·8	1,950	963·2
810·8	370	34·68	151·5	1,980	993·0
789·5	380	36·58	150·0	2,000	1,013
769·2	390	38·53	142·9	2,100	1,117
750·0	400	40·53	136·4	2,200	1,226
731·7	410	42·58	130·4	2,300	1,340
714·3	420	44·68	125·0	2,400	1,459
697·7	430	46·84	120·0	2,500	1,583
681·8	440	49·04	115·4	2,600	1,712
666·7	450	51·29	111·1	2,700	1,847
652·2	460	53·60	107·1	2,800	1,986
638·3	470	55·95	103·4	2,900	2,130
625·0	480	58·36	100·0	3,000	2,280
612·2	490	60·82	96·77	3,100	2,434
600·0	500	63·33	93·75	3,200	2,594
588·2	510	65·88	90·91	3,300	2,758
576·9	520	68·49	88·24	3,400	2,928
566·0	530	71·15	85·71	3,500	3,103
555·6	540	73·86	80·00	3,750	3,562
545·4	550	76·62	78·95	3,800	3,658
535·7	560	79·44	76·92	3,900	3,853
526·3	570	82·30	75·00	4,000	4,053
517·1	580	85·21	71·34	4,200	4,468
508·5	590	88·17	70·59	4,250	4,575
500·0	600	91·19	66·67	4,500	5,129
461·5	650	107·0	63·83	4,700	5,595
428·6	700	124·1	63·16	4,750	5,715
400·0	750	142·5	60·00	5,000	6,333
375·0	800	162·1	54·54	5,500	7,662
352·9	850	183·0	50·00	6,000	9,119
333·3	900	205·2	46·15	6,500	10,702
315·8	950	228·6	42·86	7,000	12,412
300·0	1,000	253·3	40·00	7,500	14,248
285·7	1,050	279·3	37·50	8,000	16,211
272·7	1,100	306·5	35·29	8,500	18,301
260·9	1,150	335·0	34·29	8,750	19,393
250·0	1,200	364·8	33·33	9,000	20,517
241·9	1,240	389·5	31·58	9,500	22,861
240·0	1,250	395·8	30·00	10,000	25,330
232·6	1,290	421·5	27·27	11,000	30,650
230·8	1,300	428·1	25·00	12,000	36,476
222·2	1,350	461·6	24·19	12,400	38,948
214·3	1,400	496·5	23·08	13,000	42,808
212·8	1,410	503·6	21·43	14,000	49,647
206·9	1,450	532·6	20·00	15,000	56,993
202·7	1,480	554·8	18·75	16,000	64,846
200·0	1,500	569·9	17·65	17,000	73,204
193·5	1,550	608·6	16·67	18,000	82,070
187·5	1,600	648·5	15·79	19,000	91,442
181·8	1,650	689·6	15·00	20,000	101,321
176·5	1,700	732·0	12·00	25,000	158,314
172·4	1,740	766·9	10·00	30,000	227,973
171·4	1,750	775·7			

Appendix "B."

Table II.

Frequency (f) in megacycles per second Mc/s.	Wave-length (λ) in metres (below one metre in cm.)	Frequency (f) in megacycles per second Mc/s.	Wave-length (λ) in metres (below one metre in cm.)	Frequency (f) in megacycles per second Mc/s.	Wave-length (λ) in metres (below one metre in cm.)
30·0	10·0	45·4	6·6	93·8	3·2
30·3	9·9	46·2	6·5	96·8	3·1
30·6	9·8	46·9	6·4	100·0	3·0
30·9	9·7	47·6	6·3	103·0	2·9
31·3	9·6	48·4	6·2	107·0	2·8
31·6	9·5	49·2	6·1	111·0	2·7
31·9	9·4	50·0	6·0	115·0	2·6
32·3	9·3	50·8	5·9	120·0	2·5
32·6	9·2	51·8	5·8	125·0	2·4
33·0	9·1	52·7	5·7	130·0	2·3
33·3	9·0	53·6	5·6	136·0	2·2
33·7	8·9	54·6	5·5	143·0	2·1
34·1	8·8	55·6	5·4	150·0	2·0
34·5	8·7	56·6	5·3	158·0	1·9
34·9	8·6	57·7	5·2	167·0	1·8
35·3	8·5	58·8	5·1	176·0	1·7
35·7	8·4	60·0	5·0	187·0	1·6
36·1	8·3	61·2	4·9	200·0	1·5
36·6	8·2	62·5	4·8	214·0	1·4
37·0	8·1	63·8	4·7	231·0	1·3
37·5	8·0	65·2	4·6	250·0	1·2
38·0	7·9	66·6	4·5	273·0	1·1
38·5	7·8	68·2	4·4	300·0	1·0
39·0	7·7	69·8	4·3	(f) in Mc/s.	(λ) in cms.
39·5	7·6	71·5	4·2	333·0	90
40·0	7·5	73·2	4·1	375·0	80
40·6	7·4	75·0	4·0	429·0	70
41·1	7·3	77·0	3·9	500·0	60
41·7	7·2	79·0	3·8	600·0	50
42·3	7·1	81·1	3·7	750·0	40
42·9	7·0	83·4	3·6	1,000	30
43·5	6·9	85·8	3·5	1,500	20
44·1	6·8	88·2	3·4	3,000	10
44·8	6·7	91·0	3·3	30,000	1

APPENDIX "C."

RESUSCITATION FROM APPARENT DEATH BY ELECTRIC SHOCK.

The urgent necessity for prompt and persistent efforts at resuscitation of victims of accidental shocks by electricity is very well emphasised by the successful results in the instances recorded. In order that the task may not be undertaken in a half-hearted manner, it must be appreciated that accidental shocks seldom result in death unless the victim is left unaided too long, or efforts at resuscitation are stopped too early.

The result of an electric shock on the body is to affect the nervous system, and the muscles contract involuntarily. Those controlling the heart and breathing action are stopped, and the patient appears dead. It is impossible to dogmatise on the actual voltage required to give a fatal shock. The electrical resistance of the human body is approximately 30,000 ohms when the skin is perfectly dry, but may be as low as 200-300 ohms if the skin is wet. In that case a voltage as low as 100 has proved fatal. Under normal circumstances any voltage higher than 250 should be considered dangerous (*cf.* A.31).

In the majority of instances the shock is only sufficient to suspend animation temporarily, owing to the momentary and imperfect contact of the conductors, and also on account of the resistance of the body submitted to the influence of the current. It must be appreciated that the body under the conditions of accidental shocks seldom receives the full force of the current in the circuit, but only a shunt current, which may represent a very insignificant part of the whole.

When an accident occurs, the following rules should be promptly executed with care and deliberation pending the arrival of a doctor :—

(1) Remove the body at once from the circuit by breaking contact with the conductors. This may be accomplished by using a dry stick of wood, which is a non-conductor, to roll the body over to one side, or brush aside a wire, if that is conveying the current. When a stick is not at hand, any dry piece of clothing may be utilised to protect the hand in seizing the body of the victim, unless rubber gloves are available. If the body is in contact with the earth, the coat tails of the victim, or any loose or detached piece of clothing, may be seized with impunity to draw him away from the conductor. When this has been accomplished, proceed according to (2). The object to be attained is to make the subject breathe, and if this can be accomplished and continued, he can be saved.

(2) Lay the man on the ground, face downwards. Turn his head on one side. No time should be lost by removing or loosening clothes. Begin artificial respiration by the Schaffer method, at once, as follows :—Tell one of the bystanders to prepare some sort of pad like a folded coat and slip it in under the patient's body just above his waist ; but do not wait for this. You will probably have performed several movements of respiration before the pad is ready and have thus gained all-valuable time. Kneel by the patient's side, or across his body facing his head. Spread your hands out flat on his back at his lowest ribs, one on each side, the thumbs being close to and parallel with the spine. Press gradually and slowly for about three seconds by leaning forward on to your hands. Use no violence. Relax the pressure by falling back into your original upright kneeling position for two seconds without lifting your hands from the patient. The process of artificial respiration consists in repeating this swaying motion backwards and forwards about 12 to 15 times a minute. The efforts to restore breathing must be carried out with perseverance, as in some cases it has been restored after a long period of apparent death. Keep the patient warm.

(3) The dashing of cold water into the face will sometimes produce a gasp and start breathing, which should then be continued as directed above. If this is not successful the spine should be rubbed vigorously with a piece of ice. It is both useless and unwise to attempt to administer stimulants to the victim in the usual manner by pouring them down his throat.

(4) If the patient has been burned, oil should not be used. Sterilised wool should be applied. In the case of burns resulting from accidental shock, where the respiratory system has not been affected, a patient may appear perfectly well when once his burns are dressed. It is important to realise that in all cases of electric shock there is danger of hyperstatic pneumonia setting in. The patient should therefore be kept warm and quiet for at least a day following the accident.

APPENDIX "D."

W/T TEXT BOOKS, WORKS OF REFERENCE AND JOURNALS.

The following is a list of selected works which may serve as a guide to any officer or rating who wishes to study other aspects of Radio Engineering, or who desires to read other books in conjunction with the Admiralty Handbook.

The selection is in no way an exhaustive one, and works of a highly specialised nature have been purposely excluded.

TEXT BOOKS.

"Radio Engineering." By F. E. Terman. (McGraw-Hill. 1937.—30*s.*)

"Principles of Radio Engineering." By R. S. Glasgow. (McGraw-Hill. 1936.—24*s.*)

"Modern Radio Communication." Vols. I & II. By J. H. Reyner. (Pitman. 1935.—5*s.* and 7*s.* 6*d.*, from the publisher.)

"Principles of Radio Communication." By J. Morecroft. (Chapman & Hall. 1933.—37*s.* 6*d.*)

"Short Wave Wireless Communication." By A. W. Ladner & C. R. Stoner. (Chapman & Hall. 1936 —15*s.*)

"Foundations of Wireless." By A. L. M. Sowerby. (Iliffe & Sons. 1936.—4*s.* 6*d.*)

"Wireless Receivers." By C. W. Oatley. (Methuen Co. 1932.—2*s.* 6*d.*)

"Wireless Engineering." By L. S. Palmer. (Longmans Green & Co. 1936.—21*s.*)

"Theory of Radio Communication." By the Post Office Engineering Department. (H.M.S.O. 1934.—7*s.*)

ELEMENTARY TEXT BOOKS.

"Wireless—Its Principles and Practice." By R. W. Hutchinson. (London University Tutorial Press. 1935.—3*s.* 6*d.*)

"Physical Principles of Wireless." By J. A. Ratcliffe. (Methuen Co. 1929—2*s.* 6*d.*)

"The Outline of Wireless." By R. Stranger. (Newnes, 1931.—8*s.* 6*d.*)

"Tuning in Without Tears." By Frank Boyce. (Pitman. 1936.—2*s.* 6*d.*)

TEXT BOOKS OF AMATEUR RADIO SOCIETIES.

"A Guide to Amateur Radio." (By Radio Society of Great Britain. Yearly.—6*d.*)

"The Radio Amateur's Handbook." (American Radio Relay League. Yearly.—5*s.* 6*d.*)

HISTORICAL WORKS.

"Radio Communication—History & Development." The Science Museum Handbook. (H.M.S.O. 1934.—2*s.* 6*d.*)

NUMERICAL EXERCISES IN RADIO ENGINEERING.

"Problems in Radio Engineering." By E. T. A. Rapson. (Pitman. 1935.—3*s.* 6*d.*)

"Classified Examples in Electrical Engineering—Vol. II. Alternating Current." By S. G. Monk. (Pitman. 1933.—3*s.* 6*d.*)

MORE ADVANCED TEXT BOOKS; WORKS OF REFERENCE.

"Radio Engineering Handbook." By R. Henney. (McGraw-Hill. 1935.—30*s.*)

"Wireless." By L. B. Turner. (Cambridge University Press. 1931.—25*s.*)

"Phenomena in High Frequency Systems." By A. Hund. (McGraw-Hill. 1936.—30*s.*)

"Measurements in Radio Engineering." By F. E. Terman. (McGraw-Hill. 1935.—24*s.*)

"Thermionic Vacuum Tubes." By E. V. Appleton. (Methuen Co. 1931.—3*s.*)

"Atmospheric Electricity." By B. F. J. Schonland. (Methuen Co. 1932.—2*s.* 6*d.*)

"Electromagnetic Waves." By F. W. G. White. (Methuen Co. 1934.—3*s.*)

APPENDIX "D."

JOURNALS.

"Wireless World." (Iliffe & Sons. Weekly.—*4d.*)

"Wireless Engineering." (Iliffe & Sons. Monthly.—*2s. 6d.*)

"Proceedings of the Institute of Radio Engineers." (New York.)

"Journal of the Institution of Electrical Engineers." (Published Monthly.—*10s. 6d.*)

"Electronics." (Published Monthly. New York.)

"The Post Office Electrical Engineers' Journal." (Published by the Electrical Review Quarterly.—*1s.*)

APPENDIX "E."

THE SIGN OF MUTUAL INDUCTANCE.

(*Cf.* Vol. I, paragraph 161, and Vol. II, Section K.4.)

In the mathematical analysis of the tuned anode oscillator, and other circuits, some confusion has arisen as to the sign to be attributed to the grid oscillatory voltage. If the circuit is regarded as a mathematical exercise, the object of which is to find the conditions for oscillation, at the beginning nothing is known about the directions of the windings of the coils, or which end of the grid coil is to be connected to the grid. Under those conditions it may be assumed that the grid oscillatory voltage is given by either $+j\omega MI_L$, or $-j\omega MI_L$. With reference to K.6, if the first assumption is made, then if M is positive we obtain a vector diagram as shown in Fig. 3 of Section "K," and the circuit will oscillate. If the second assumption is made (*i.e.*, $-j\omega MI_L$), V_g on the vector diagram will be reversed and the circuit will not oscillate.

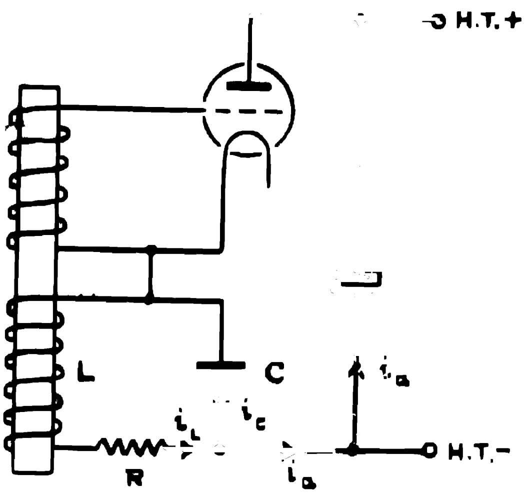


FIG. 1.

In mathematical calculations the phases of the vectors may be correctly given by saying that M itself is negative. Since some early writers wrote $V_g = -j\omega MI_L$, there has tended to arise an idea that M must be negative for a tuned anode circuit to oscillate, irrespective of what convention is used for positive or negative M in other types of circuit, not involving valve oscillators. This, of course, is entirely untrue.

In practice we are only interested in the phase of the grid input voltage, with relation to the current in the inductance in the anode circuit, and we can reverse this phase by reversing the connections to the grid coil. Thus, so far as the valve oscillator itself is concerned, we can make the voltage applied to the grid either $+j\omega MI_L$ or $-j\omega MI_L$, by simply reversing the connections.

When it is desired to specify the directions of the windings of coils, and the method of connecting them up, in practice it is found that if the two coils are wound on the same former and joined up like one continuous coil, then—

(*a*) the grid should be connected to one free end and the anode to the other; and

(*b*) the cathode should be connected to the common point.

Under these conditions the oscillatory grid voltage will be in the right phase for self-oscillation and given by $+j\omega MI_L$, where M is positive by the convention defined in Vol. I.

Fig. 1 represents an alternative way of drawing the diagram for a tuned anode oscillator. It has the advantage of emphasising a common point in the grid and anode circuits.

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